

THIRD EDITION

Modeling and Analysis of **DYNAMIC SYSTEMS**

Ramin S. Esfandiari
Bei Lu



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Third Edition



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By

Ramin S. Esfandiari

Bei Lu



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*To my wife Haleh,
my sisters Mandana and Roxana,
and my parents to whom I owe it all.*

Ramin S. Esfandiari

*To my husband Qifu,
my daughters Meixin and Meiling,
and my parents.*

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Preface

As with the earlier editions of this book, the primary goal of this new edition is to provide the reader with a thorough knowledge of mathematical modeling and analysis of dynamic systems. MATLAB, Simulink, and Simscape are introduced at the outset and are utilized throughout the book to perform symbolic, graphical, numerical, and simulation tasks. The textbook, written at the junior level, meticulously covers techniques for modeling dynamic systems, methods of response analysis, and an introduction to vibration and control systems.

This book consists of 10 chapters and 2 appendices. [Chapter 1](#) provides an introduction of MATLAB, Simulink, and Simscape to the reader. The essential mathematical background is covered in [Chapters 2](#) and [3](#). Different forms of system model representation (state-space form, transfer function, input–output equation, block diagram, etc.) and linearization are discussed in [Chapter 4](#). Each topic is also handled using MATLAB, and block diagrams are constructed and analyzed using Simulink.

[Chapter 5](#) discusses translational, rotational, and mixed mechanical systems. Free-body diagram approach is emphasized in the derivation of the systems' equations of motion. Electrical and electromechanical systems are covered in [Chapter 6](#) and operational amplifiers and impedance methods are also included in [Chapter 6](#). [Chapter 7](#) presents pneumatic, liquid-level, and thermal systems. Modeling and analysis of dynamic systems ranging from mechanical to thermal using Simulink and Simscape are fully integrated in [Chapters 5](#) through [7](#).

Time-domain and frequency-domain analyses of dynamic systems are introduced in [Chapter 8](#). Time-domain analysis entails transient response of first-, second-, and higher-order systems. The sinusoidal transfer function (frequency response function) is introduced and utilized in obtaining the system's frequency response as well as Bode diagram. Analytical solution of the state equation is also included in this chapter. MATLAB and Simulink play significant roles in determining and simulating system response and are used throughout the chapter.

[Chapter 9](#) presents an introduction to vibrations and includes free and forced vibrations of single and multiple degrees-of-freedom systems, vibration suppression including vibration isolators and absorbers, modal analysis, and vibration testing. Some applications of vibrations are also included: logarithmic decrement for experimental determination of the damping ratio, rotating unbalance, and harmonic base excitation.

[Chapter 10](#) gives an introduction to control systems analysis and design in the time and frequency domains. Basic concepts and terminology are presented first, followed by stability analysis, system identification, types of control, root-locus analysis, Bode plot, and full-state feedback. These techniques are subsequently implemented using MATLAB, Simulink, and Simscape.

Appendices

[Appendix A](#) includes a summary of systems of units and conversion tables. [Appendix B](#) contains useful formulas such as trigonometric identities and integrals.

Examples and Exercises

Each covered topic is followed by at least one example for a better understanding of the subject matter. More complex topics are accompanied by multiple, thoroughly worked-out examples. Each section of each chapter is followed by several exercises so that the reader can immediately apply the ideas that are just learned. The end-of-chapter review exercises help to learn how a combination of different ideas can be used to analyze a problem.

Chapter Summaries

Chapter summaries provide concise reviews of the key aspects of each chapter.

Instructor's Solutions Manual

A solutions manual, featuring complete solution details of all exercises, is prepared by the authors and will be available to instructors adopting the book.

New to This Edition

The main new features of this edition are as follows:

1. System identification is introduced and utilized to construct models of dynamic systems from observed input–output data.
2. The end-of-chapter exercises ([Chapters 4](#) through [10](#)) now include case studies that involve the application of most of the concepts and methods covered in the chapter to real-world problems.

3. At least 50% of the examples and exercises throughout the book are either new or have been dramatically revised. The problem sets also include more challenging exercises.

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1

Introduction to MATLAB, Simulink, and Simscape

This chapter introduces the fundamental features of MATLAB®, Simulink®, and Simscape™ that are pertinent to modeling and analysis of dynamic systems. These include the descriptions and applications of several commonly used built-in functions (commands) in MATLAB and the basics of building block diagrams for the purpose of simulation of dynamic systems by using Simulink and Simscape. MATLAB, Simulink, and Simscape are fully integrated throughout the book, and the fundamental features and capabilities presented in this chapter will play an important role in better understanding the more advanced applications in the subsequent chapters.

1.1 MATLAB Command Window and Command Prompt

Once a MATLAB session is opened, commands can be entered at the MATLAB command prompt ">>"; see [Figure 1.1](#). For example, $\sqrt{\cos(\frac{1}{6}\pi)}$ may be calculated as

```
>> sqrt(cos(pi/6))
ans =
    0.9306
```

The outcome of a calculation can be stored under a variable name and suppressed by using a semicolon at the end of the statement:

```
>> sc = sqrt(cos(pi/6));
```

Commands such as `sqrt` (square root) and `cos` (cosine of an angle in radians) are MATLAB built-in functions. Each of these functions is accompanied by a brief but sufficient description through the `help` command. For example,

```
>> help sqrt
sqrt    Square root.
        sqrt(X) is the square root of the elements of X. Complex
        results are produced if X is not positive.
        See also sqrtm, realsqrt, hypot.
        Overloaded methods:
            sym/sqrt
        Reference page in Help browser
            doc sqrt
```

For a variable "x," other elementary functions include `abs(x)` for $|x|$, `sin(x)` for $\sin x$, `log(x)` for $\ln x$, `log10(x)` for $\log x$, `exp(x)` for e^x , and many more. Descriptions of these functions are available through the `help` command.

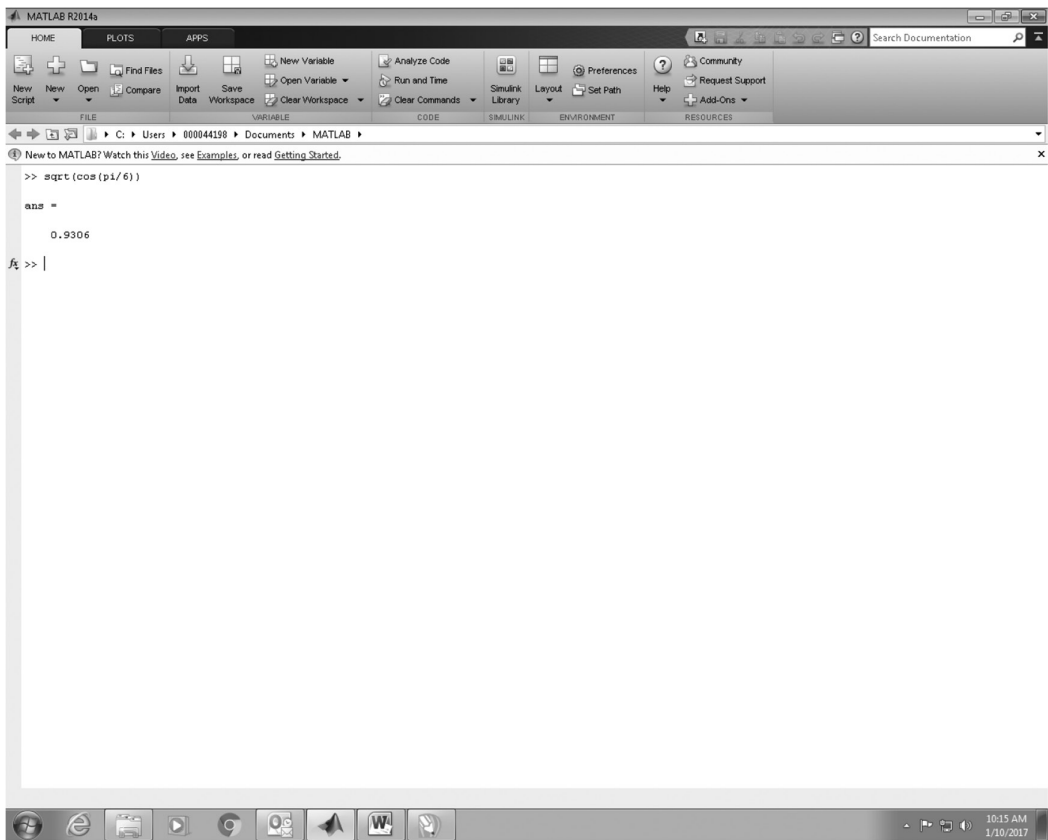


FIGURE 1.1
Screen capture of a MATLAB session.

1.2 Vectors and Matrices

Vectors can be created and handled in various ways in MATLAB. A row vector is created as

```
>> v = [-2 3 0 4 8];
```

Commas may be used instead of spaces between elements. Column vectors are created when semicolons separate the vector components.

```
>> w = [-2;3;0;4;8];
```

The length of a (row or column) vector is determined by using the `length` command:

```
>> length(v)
ans =
    5
```

The size of a vector is determined by the `size` command. For the column vector `w` defined earlier,

```
>> size(w)
ans =
     5     1
```

Arrays of evenly spaced numbers can be created more effectively as follows. For example, a row vector whose first element is 0 and last element is 20, with a spacing of 4, is created as

```
>> v = [0:4:20]      or      >> v = 0:4:20
v =
     0     4     8    12    16    20
```

To create a column vector with the same properties:

```
>> w = [0:4:20] '      or      >> w = v'
w =
     0
     4
     8
    12
    16
    20
```

Specific component(s) of a vector can be easily retrieved. For example, the fourth component of vector `v` or `w` is retrieved by typing

```
>> v(4)      or      >> w(4)
ans =
    12
```

A group of components may also be retrieved. For example, the last four components of `v` are recovered as

```
>> v(end-3:end)
ans =
     8    12    16    20
```

1.2.1 Linspace

Vectors with equally spaced elements may also be created by using the `linspace` command.

```
>> x = linspace(1,10,5) % Five equally spaced points between 1 and 10
x =
    1.0000    3.2500    5.5000    7.7500   10.0000
```

The default value for the number of points is 100. Therefore, `x = linspace(1,10)` generates 100 equally spaced points between 1 and 10.

1.2.2 Matrices

A matrix is created by using brackets enclosing all of its elements; rows are separated by a semicolon.

```
>> A = [2 0 -4;1 3 5;-3 2 6]
```

```
A =
     2     0    -4
     1     3     5
    -3     2     6
```

A matrix entry is accessed by using the row and column number of the location of that entry.

```
>> A(2,3) % Entry at the intersection of the second row and third column
ans =
     5
```

An entire row or column of a matrix is accessed by using a colon.

```
>> Row_3 = A(3,:) % Third row of A
Row_3 =
    -3     2     6

>> Col_2 = A(:,2) % Second column of A
Col_2 =
     0
     3
     2
```

To replace an entire row of matrix A by a given vector v, we proceed as follows:

```
>> v = [3 1 7];
>> A_new = A; % Pre-allocate the new matrix
>> A_new(2,:) = v % Replace the second row with v
A_new =
     2     0    -4
     3     1     7
    -3     2     6
```

The $m \times n$ zero matrix is created by using `zeros(m,n)`; for instance, the 2×3 zero matrix:

```
>> Z = zeros(2,3);
```

The $m \times n$ zero matrix is commonly used for pre-allocation of matrices to save memory space. In the zero matrix Z defined earlier, any entry can be altered, whereas others remain unchanged.

```
>> Z(1,3) = 3; Z(2,1) = -1
```

```
Z =
     0     0     3
    -1     0     0
```

Size of a matrix is determined by using the `size` command:

```
>> size(Z)
```

```
ans =
     2     3
```

The $n \times n$ identity matrix is created by `eye(n)`. Matrix operations can be easily performed in MATLAB. If the sizes are not compatible, or the operations are not defined, MATLAB returns an error message to that effect.

```
>> A = [1 -1 1; 2 0 4]; B = [-3 1; 5 2; 0 6]; % A is 2-by-3, B is 3-by-2
>> C = A*B % Operation is valid
```

```
C =
    -8     5
    -6    26
```

1.2.3 Determinant, Transpose, and Inverse

The determinant of an $n \times n$ matrix is calculated by the `det` command. The transpose of a matrix A is found as

```
>> A = [0 1 0; 0 0 1; -1 -2 -3]; At = A.'
```

```
At =
     0     0    -1
     1     0    -2
     0     1    -3
```

The inverse of a nonsingular (non-zero determinant) matrix is handled by the `inv` command.

1.2.4 Slash Operators

There are two slash operators in MATLAB: backslash (`\`) and slash (`/`).

```
>> help \
```

```
\ Backslash or left matrix divide.
```

```
A\B is the matrix division of A into B, which is roughly the
```


same as $\text{INV}(A)*B$, except it is computed in a different way. If A is an N -by- N matrix and B is a column vector with N components, or a matrix with several such columns, then $X = A \setminus B$ is the solution to the equation $A*X = B$. A warning message is printed if A is badly scaled or nearly singular. $A \setminus \text{EYE}(\text{SIZE}(A))$ produces the inverse of A .

The solution of a linear system of algebraic equations $Ax = b$ is ordinarily found via $x = A^{-1}b$. However, instead of performing $x = \text{inv}(A)*b$, it is more efficient to find it via $x = A \setminus b$. The description of the slash (/) operator is given in the following.

```
>> help /

/   Slash or right matrix divide.

A/B is the matrix division of B into A, which is roughly the
same as A*INV(B), except it is computed in a different way.
More precisely, A/B = (B'\A')'. See MLDIVIDE for details.
```

1.2.5 Element-by-Element Operations

Element-by-element operations are summarized in [Table 1.1](#). These are used when operations are performed on each element of a vector or matrix.

For example, consider the vector

```
>> v = [1 -2 3];
```

To raise each element of v to power of 2,

```
>> v.^2           % If v^2 is used, an error message is returned by MATLAB

ans =

     1     4     9
```

Now, consider $(2+v)/(1+v)$, where vector v is defined previously. This fraction is to be evaluated for each component of v :

```
>> (2.+v) ./ (1.+v)

ans =

    1.5000         0    1.2500
```

TABLE 1.1
Element-by-Element Operations

MATLAB Symbol	Description
.*	Multiplication
./	(Right) Division
.^	Exponentiation

Arrays of the same size can be handled in the element-by-element operation.

```
>> v = [1;2;-3]; w = [4;-2;2];
>> v.*w
ans =
     4
    -4
    -6
```

1.3 Symbolic Math Toolbox

The symbolic math toolbox allows for the manipulation of symbols to perform operations symbolically. Symbolic variables are created by using the `syms` command. For example, the function $g = e^{-2a} + \frac{1}{2}\sin 3b$ can be defined symbolically as follows:

```
>> syms a b
>> g = exp(-2*a)+sin(3*b)/2
g =
exp(-2*a) + sin(3*b)/2
```

This function may also be defined symbolically via

```
>> g = sym('exp(-2*a)+sin(3*b)/2');
```

In this situation, variables `a` and `b` do not need to be declared symbols, as this is handled automatically by `sym` in the definition of `g`. Assignment of specific values to variables will not be taken into account when using `sym`. Instead, we can use the `subs` command at this stage:

```
>> a = 1.6; b = 2.4; g1 = subs(g)
g1 =
exp(-16/5) + sin(36/5)/2
```

In symbolic expressions, numbers are converted to the ratio of two integers, as it is observed here as well. For decimal representation of numbers, we use the `double` command to convert to double precision.

```
>> a = 1.6; b = 2.4; g1 = double(subs(g))
g1 =
    0.4376
```

1.3.1 Anonymous Functions

An anonymous function offers a way to create a function for simple expressions. Anonymous functions can only contain one expression and cannot return more than one output variable. The generic form is

```
My_function = @(arguments)(expression)
```

As an example, the function $Q = \sqrt{2 + e^{-ax/3}}$ may be created as

```
>> Q = @(a,x) (sqrt(2+exp(-a*x/3)));
```

This creates $Q(a,x)$, which may then be evaluated for specific values of a and x . For example,

```
>> Q(1,-2)
ans =
    1.9869
```

1.3.2 MATLAB Function

The built-in `matlabFunction` allows us to generate a MATLAB file or an anonymous function from sym object. The generic form $G = \text{matlabFunction}(F)$ converts the symbolic expression or function F to a MATLAB function with the handle G .

The function $Q = \sqrt{2 + e^{-ax/3}}$, considered earlier, can now be created via

```
>> syms a x
>> Q = matlabFunction(sqrt(2+exp(-a*x/3)))
Q =
    @(a,x) sqrt(exp(a.*x.*(-1.0./3.0))+2.0) % Variables arranged in
    alphabetical order

>> Q(1,-2)
ans =
    1.9869
```

As expected, this agrees with the earlier result using the anonymous function. Note that if the desired order of variables is not specified by the user, MATLAB will list them in an alphabetical order. In the above-mentioned example, omitting the list of variables would still result in $Q(a, x)$. However, if $Q(x, a)$ is desired, the 'vars' option is utilized as follows:

```
>> Q = matlabFunction(sqrt(2+exp(-a*x/3)), 'vars', [x a])
Q =
    @(x,a) sqrt(exp(a.*x.*(-1.0./3.0))+2.0)
```

1.3.3 Differentiation

In order to find the derivative of a function with respect to any of its variables, the function must be defined symbolically. For example, consider $f(t) = t^2 - \frac{2}{3} \cos t$, a function of a single variable. To determine df/dt , we proceed as follows:

```
>> f = sym('t^2-2*cos(t)/3');
>> dfdt = diff(f)
dfdt =
    2*t + (2*sin(t))/3
```

The second derivative d^2f/dt^2 is found as

```
>> dfdt2 = diff(f,2)
dfdt2 =
    (2*cos(t))/3 + 2
```

The symbolic derivatives can be converted to MATLAB functions for convenient evaluation. For example, to evaluate df/dt when $t = 2.14$,

```
>> f = sym('t^2-2*cos(t)/3');           % Define function symbolically
>> fd = matlabFunction(diff(f));         % Convert derivative to MATLAB function

>> fd(2.14)                             % Evaluate the derivative
ans =

    4.8416
```

1.3.4 Integration

Indefinite and definite integrals are calculated symbolically via the `int` command.

```
int      Integrate
int(S) is the indefinite integral of S with respect to its symbolic
        variable as defined by SYMVAR. S is a SYM (matrix or scalar).
        If S is a constant, the integral is with respect to 'x'.

int(S,v) is the indefinite integral of S with respect to v. v is a
        scalar SYM.

int(S,a,b) is the definite integral of S with respect to its
        symbolic variable from a to b. a and b are each double or
        symbolic scalars.

int(S,v,a,b) is the definite integral of S with respect to v
        from a to b.
```

For example, the indefinite integral $\int (a + e^{-t/2}) dt$, where a is a parameter, is calculated as

```
>> f = sym('a+exp(-t/2)');
>> int(f)
ans =
a*t - 2*exp(-t/2)
```

The definite integral $\int_{-1}^2 (a + e^{-t/2}) dt$ is calculated as follows:

```
>> f = sym('a+exp(-t/2)');
>> syms t
>> I = int(f,t,-1,2)
I =
3*a + 2*exp(-1)*(exp(3/2) - 1)
```

To evaluate the integral when $a = 0.36$, we proceed as follows:

```
>> a = 0.36; double(subs(I))
ans =
    3.6417
```

Note that the default integration variable here is t . Thus, in the above-mentioned example, it could have been omitted to yield the correct result:

```
>> int(f,-1,2);
```

Definite integrals can also be handled by the `integral` command.

```
>> help integral
integral Numerically evaluate integral.
    Q = integral(FUN,A,B) approximates the integral of function FUN from A
    to B using global adaptive quadrature and default error tolerances.

    FUN must be a function handle. A and B can be -Inf or Inf. If both are
    finite, they can be complex.
```

The `integral` command requires that the integrand be a function handle, and limits of integration specified. Therefore, to evaluate $\int_{-1}^2 (a + e^{-t/2}) dt$ when $a = 0.36$, we proceed as follows:

```
>> f = @(a,t) (a+exp(-t/2));
>> integral(@(t)f(0.36,t),-1,2)

ans =

    3.6417
```

Of course, this agrees with the earlier result using the `int` command.

1.3.5 Differential Equations

Differential equations and initial-value problems can be solved by the `dsolve` function. For example, the solution of the differential equation $xy' + (x+1)y = 0$ is obtained as

```
>> y = dsolve('x*Dy+(x+1)*y=0','x')
y =
    (C3*exp(-x))/x % C3 is some constant
```

Note that the default independent variable in `dsolve` is t . Since in our example the independent variable was x , we needed to specify that in single quotes. The initial-value problem $\ddot{x} + 2\dot{x} + 2x = e^{-t/3}$, $x(0) = 0$, $\dot{x}(0) = 1$, with t as the independent variable, is solved as

```
>> x = dsolve('D2x+2*Dx+2*x=exp(-t/3)','x(0)=0, Dx(0)=1'); x =
simplify(x)
x =
    (exp(-t)*(9*exp((2*t)/3) - 9*cos(t) + 7*sin(t)))/13
```

The solution thus obtained is symbolic and hence must be converted into a MATLAB function before it can be evaluated for specific values of the independent variable.

1.4 Plotting

Plotting a vector of values versus another vector of values is handled by the `plot` command.

`plot` Linear plot.

`plot(X,Y)` plots vector `Y` versus vector `X`. If `X` or `Y` is a matrix, then the vector is plotted versus the rows or columns of the matrix, whichever line up. If `X` is a scalar and `Y` is a vector, disconnected line objects are created and plotted as discrete points vertically at `X`.

For example, we plot the function $x(t) = e^{-t/2}(\sin t + \cos t)$ over the interval $[0, 10]$ as follows. Note that the default number of points to be used to generate a smooth curve is 100.

```
>> t = linspace(0,10); % 100 values for 0 ≤ t ≤ 10
>> x = exp(-t/2).*(sin(t)+cos(t)); % Corresponding 100 values for x
>> plot(t,x) %Figure 1.2
```

The Figure Window can be used to edit the figure. Its functions include adding grid, adjusting the thickness of lines and curves, adding text and legend, axes titles, and much more.

1.4.1 Subplot

The built-in function `subplot` is designed to create multiple figures in tiled positions. Suppose we want to plot the function $y(x,t) = e^{-2x} \cos(t+x)$ versus $0 \leq x \leq 2$ for $t = 0, 1, 2, 3$. Let us generate the four plots and arrange them in a 2×2 formation.

```
x = linspace(0,2); t = 0:1:3;
for i = 1:4,
    for j = 1:100,
        y(j,i) = exp(-2*x(j))*cos(t(i)+x(j)); % Generate 100 values of y for each t
    end
end
```

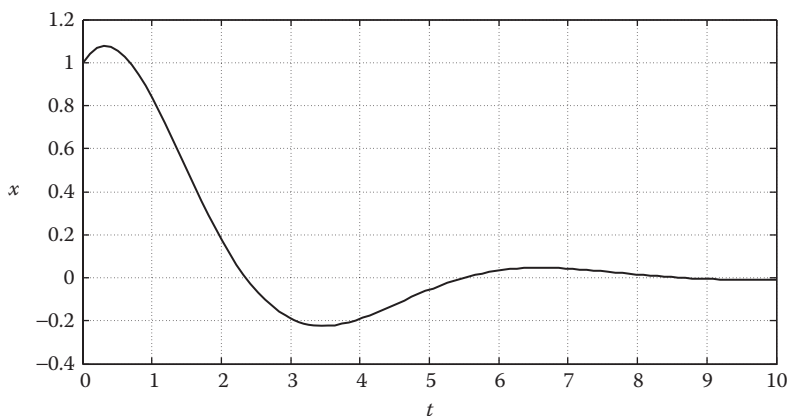


FIGURE 1.2

Plot of a function versus its variable.

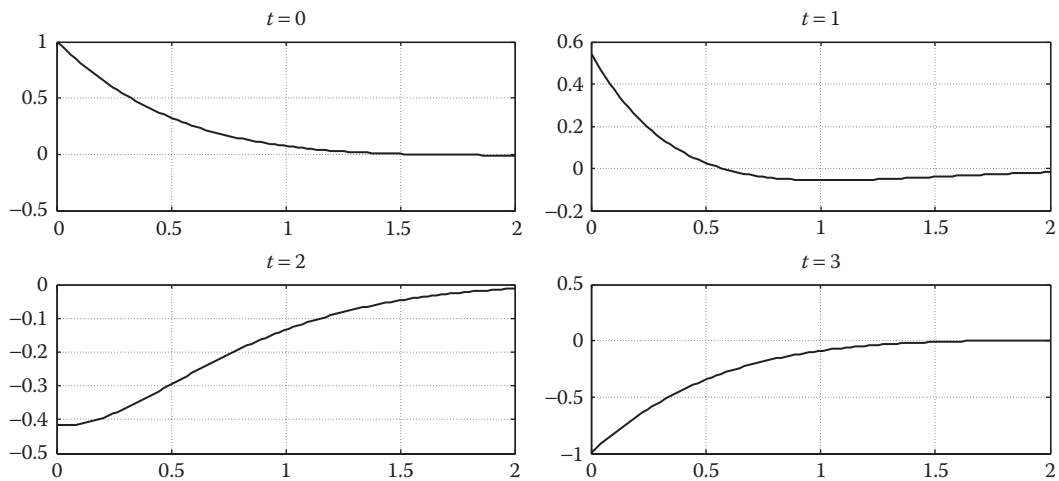


FIGURE 1.3
Multiple figures in tiled positions.

```
% Initiate Figure 1.3
subplot(2,2,1), plot(x,y(:,1)), title('t = 0')
subplot(2,2,2), plot(x,y(:,2)), title('t = 1')
subplot(2,2,3), plot(x,y(:,3)), title('t = 2')
subplot(2,2,4), plot(x,y(:,4)), title('t = 3')
```

1.4.2 Plotting Analytical Expressions

An alternative way to plot a function is to use the `ezplot` command, which plots the function without requiring data generation. As an example, consider the function $x(t) = e^{-t/2}(\sin t + \cos t)$ that we previously plotted over the interval $[0, 10]$. The plot in Figure 1.2 can be re-generated as follows:

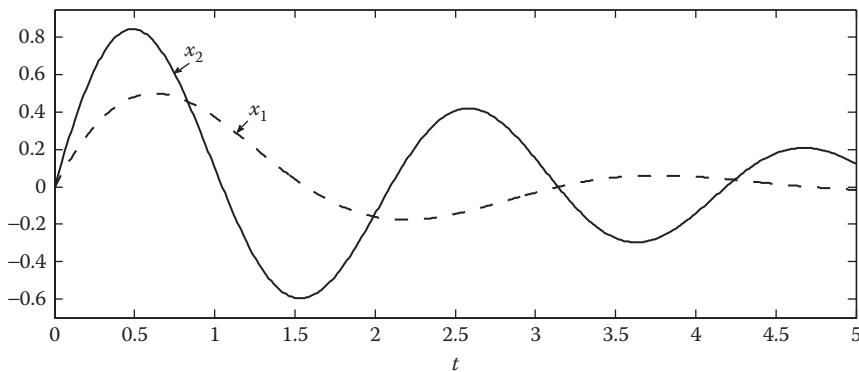
```
>> x = sym('exp(-t/2)*(cos(t)+sin(t))');
>> ezplot(x, [0,10])
```

The default variable range in `ezplot` is $[-2\pi, 2\pi]$.

1.4.3 Multiple Plots

Multiple plots can be created using `plot` (which handles vectors of values versus vectors of values) and `ezplot` (which handles analytical expressions). Suppose the two functions $x_1 = 0.8e^{-2t/3} \sin 2t$ and $x_2 = e^{-t/3} \sin 3t$ are to be plotted versus $0 \leq t \leq 5$ in the same graph.

```
>> x1 = sym('0.8*exp(-2*t/3)*sin(2*t)');
>> x2 = sym('exp(-t/3)*sin(3*t)');
>> ezplot(x1, [0,5]) % Initiate Figure 1.4
>> hold on
>> ezplot(x2, [0,5]) % Complete Figure 1.4
```

**FIGURE 1.4**

Multiple plots in a single figure.

Executing the preceding script generates a figure that does not exactly match with [Figure 1.4](#). To enable the interactive plot editing mode in the MATLAB Figure Window, click the Edit Plot button (⌘) or select Tools > Edit Plot from the main menu. If you enable plot editing mode in the MATLAB Figure Window, you can perform point-and-click editing of your graph. In this mode, you can modify the appearance of a graphic object by double-clicking the object and changing the values of its properties.

1.5 User-Defined Functions and Script Files

User-defined M file functions and scripts may be created, saved, and edited in MATLAB by using the `edit` command. For example, suppose we want to create a function called `Cyl_SV` that returns the surface area S and the volume V of a cylinder with radius r and height h . The function can be saved in a folder on the MATLAB path or in the current directory. The current directory can be viewed and/or changed using the dropdown menu at the top of the MATLAB command window. Once the current directory has been properly selected, type

```
>> edit Cyl_SV
```

A new window (Editor Window) will be opened, where the function can be created. The generic structure of a function is

```
function [output variables] = FunctionName(input variables)
% Comments
% Expressions/statements
% Calculation of all output variables
```

Our user-defined function `Cyl_SV` is specifically created as follows:

```
function [S,V] = Cyl_SV (r,h)
S = 2*pi*r^2 + 2*pi*r*h;
V = pi*r^2*h;
end
```


Executing this function for a cylinder of radius $r = 0.4$ and height $h = 1.5$ returns

```
[S,V] = Cyl_SV(0.4,1.5)
S =
    4.7752
V =
    0.7540
```

1.5.1 Creating a Script File

A script file comprises a list of commands, as if they were typed at the command line. Script files can be created in the MATLAB Editor and saved as an M file. For example, typing

```
>> edit My_script_file
```

opens the Editor Window, where the script can be created and saved under the name `My_script_file`. It is recommended that a script start with the functions `clear` and `clc`. The first one clears all the previously generated variables, and the second one clears the Command Window. Suppose we type the following lines in the Editor Window:

```
clear
clc
a = -1; b = 5;
C = sqrt(abs(3^a*cos(b)));
```

While in the Editor Window, select "Run `My_script_file.m`" under the Debug pull-down menu. This will execute the lines in the script file and return the command prompt. Simply type `C` at the prompt to see the result.

```
>> My_script_file
>> C
C =
    0.3075
```

This can also be done by highlighting the contents and selecting "Evaluate Selection." An obvious advantage of creating a script file is that it allows us to simply make changes to the contents, without having to re-type all the commands.

1.6 Simulink

Simulink is a powerful software package widely used in academia and industry for the modeling, analysis, and simulation of dynamic systems. In the modeling phase, models are built as block diagrams ([Section 4.5](#)) via a graphical user interface (GUI). Once a model is built, it can be simulated with the aid of Simulink menus or by entering commands in

MATLAB's command window. One of the greatest advantages of Simulink is the fact that it allows for the analysis and simulation of the more realistic nonlinear models, in addition to the idealized linear ones.

1.6.1 Block Library

Typing Simulink at the MATLAB command prompt opens a new window labeled Simulink Library Browser (Figure 1.5), which includes a complete block library of sinks, sources, components, and connectors. Clicking on any of the categories reveals the list of blocks it contains. For instance, clicking on Commonly Used Blocks results in the window shown in Figure 1.6.

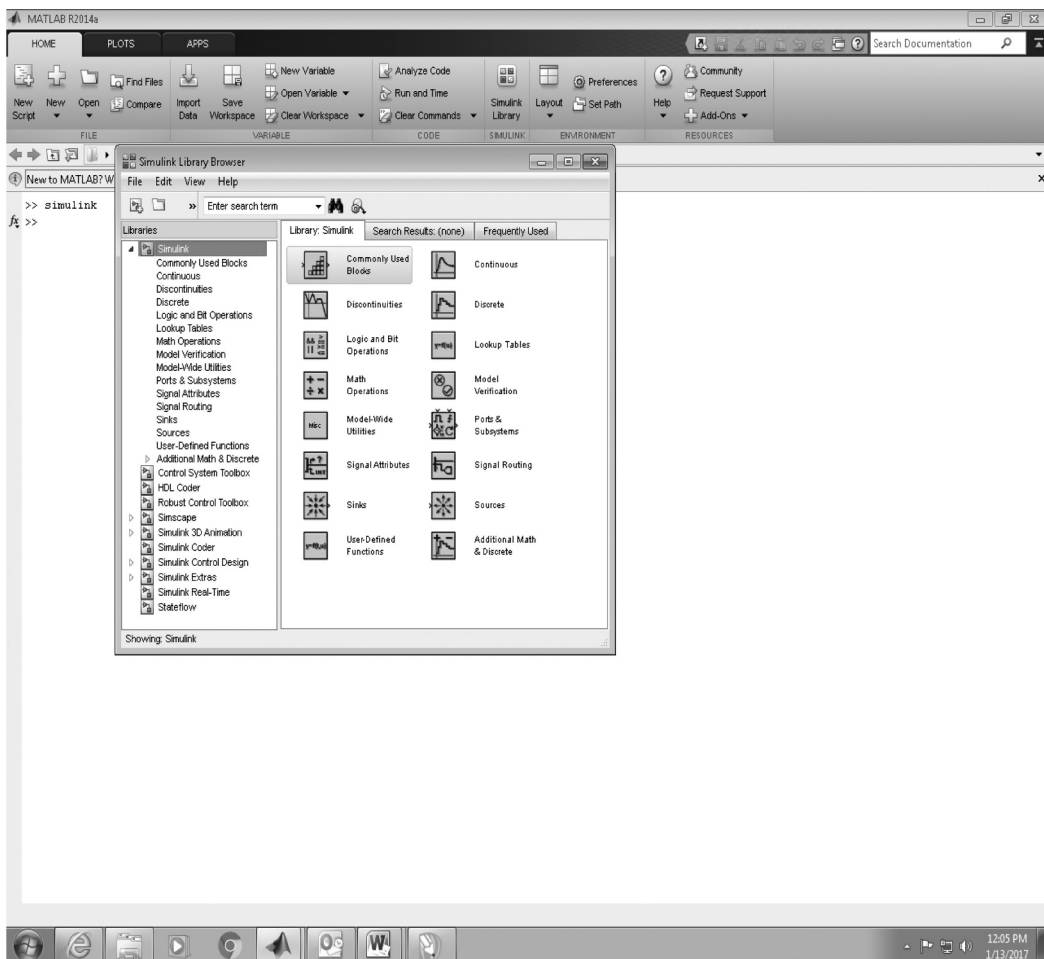


FIGURE 1.5
Simulink library browser.

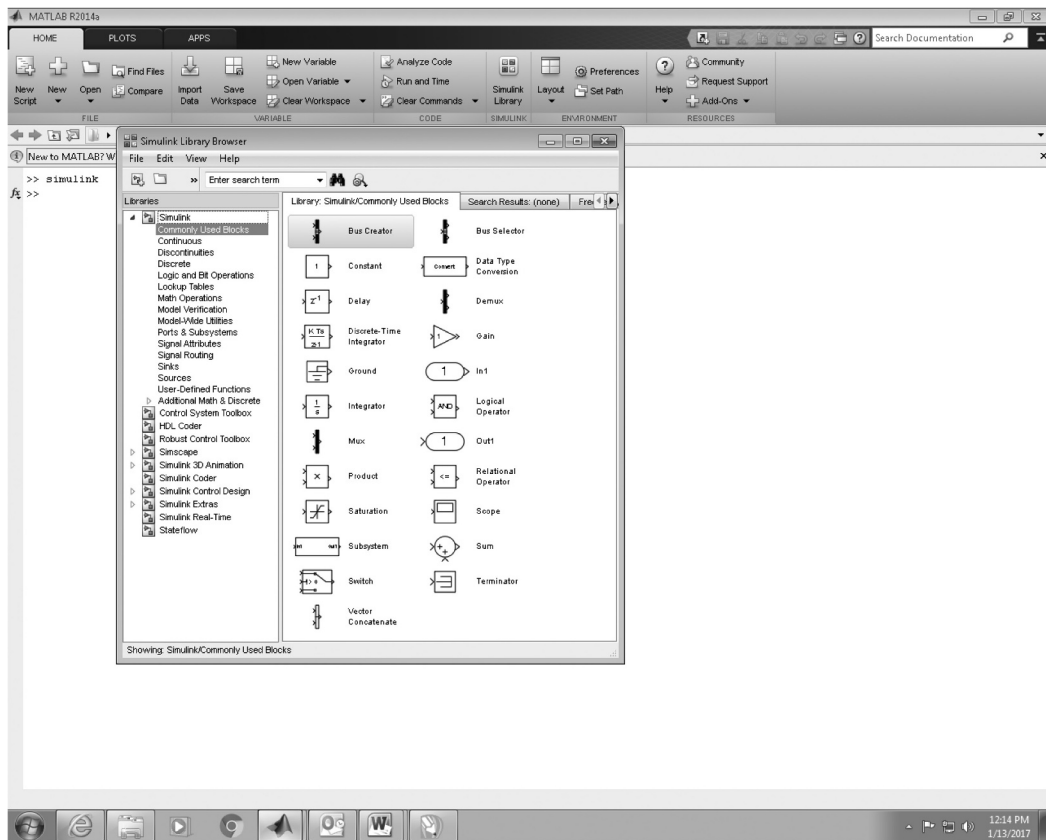


FIGURE 1.6
Commonly used blocks.

1.6.2 Building a New Model

To create a new model, select **File > New > Model** in the Simulink library window or simply press the **New Model** button (the icon underneath the **File** menu button) on the Library Browser's toolbar. This will open a new model window (Figure 1.7). Suppose we want to build a model that integrates a sine wave (amplitude 1 and frequency 1 rad/sec) and displays the result, along with the sine wave signal itself. The block diagram of this model will resemble Figure 1.8. To create this model, we will need to copy (drag and drop) the following blocks into our new model:

- The Sine Wave signal (from Sources library)
- The Integrator block (from Continuous library)
- The Mux block (from Signal Routing library)
- The Scope block (from Sinks library)

This results in what is shown in Figure 1.9. Note that double-clicking on each block reveals more detailed properties for that block. For instance, double-clicking on the Sine Wave block reveals that the amplitude has a default value of 1 and the frequency has a default

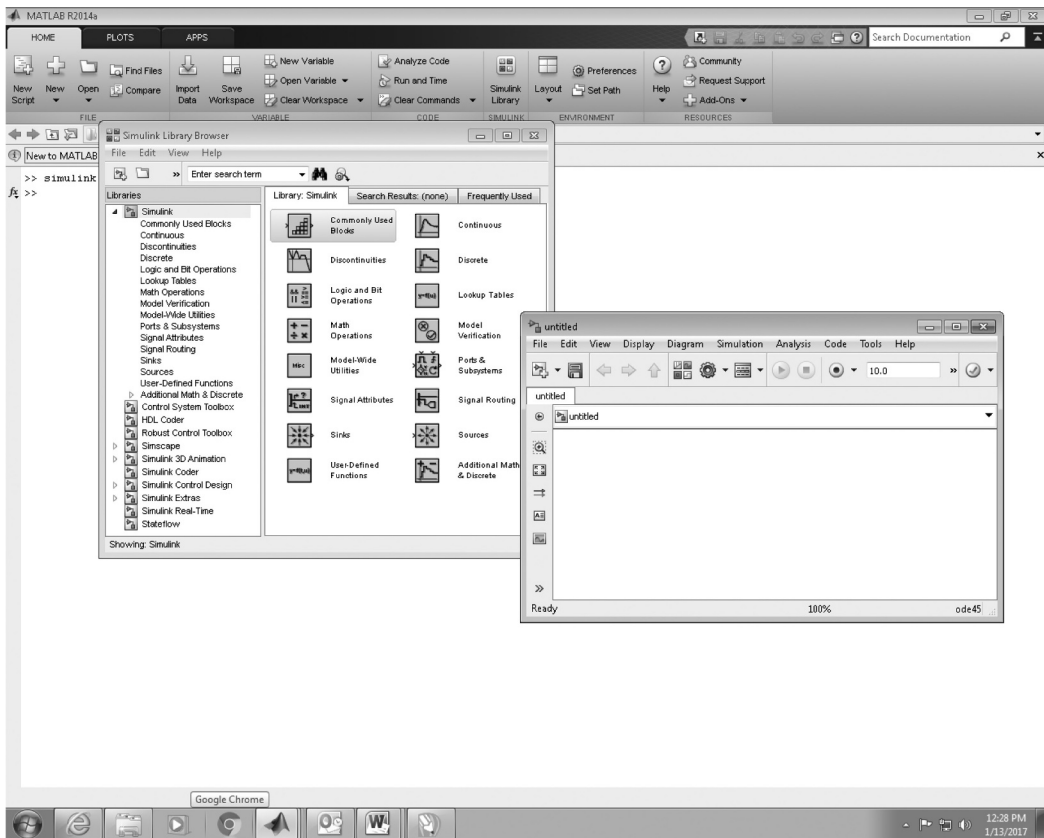


FIGURE 1.7
A new model window.

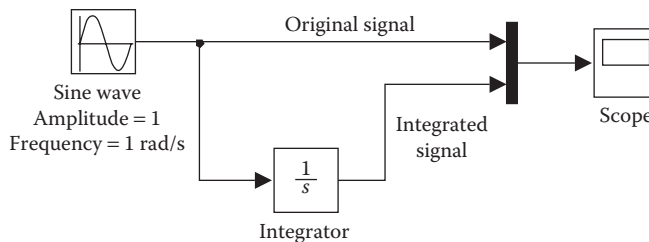
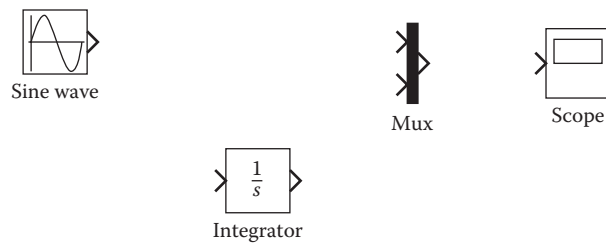


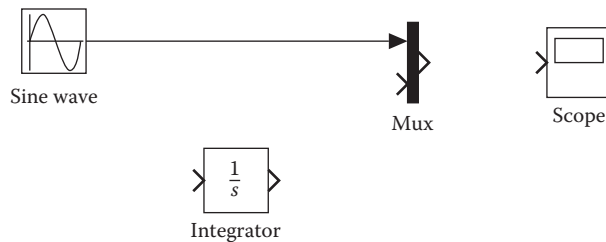
FIGURE 1.8
New Simulink model.

value of 1 rad/sec. Since these happen to be the desired parameter values for our current example, no modifications are necessary. The ">" symbol pointing out of a block is an output port. The symbol ">" pointing to a block is an input port. A signal travels through a connecting line out of an output port of a block and into an input port of another block. The port symbols disappear as soon as the blocks are connected. Let us now connect the blocks in [Figure 1.9](#). First, connect the Sine Wave block to the top input port of the

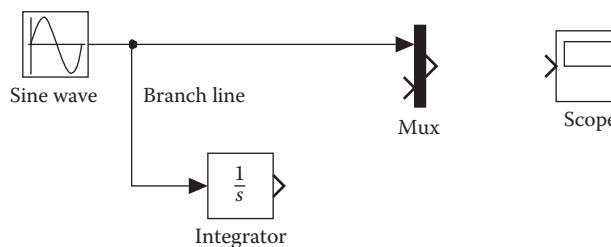
**FIGURE 1.9**

Basic blocks involved in a new model.

Mux block. Place the pointer over the output port of the Sine Wave block, and note that the cursor shape changes to crosshairs. Hold down the mouse button and move the cursor to the top input port of the Mux block. Then, release the mouse button. If the connecting line is not straight, simply drag the Sine Wave block up or down, until the ports are lined up (Figure 1.10). The signal going from the Sine Wave block to the Mux block must also go through the integrator. This may be done by means of a branch line. Place the pointer on the connecting line between the Sine Wave and Mux blocks. Press and hold down the Ctrl key. Press the mouse button, and drag the pointer to the Integrator block's input port. Then, release the mouse button. This results in Figure 1.11. Finally, connect from the output port of the Integrator to the bottom input port of the Mux block, and draw a connecting line from the output port of the Mux to the input port of the Scope. The completed block diagram will look like the original Figure 1.8.

**FIGURE 1.10**

Connecting line.

**FIGURE 1.11**

Branch line.

1.6.3 Simulation

The simulation parameters can be set by choosing Model Configuration Parameters from the Simulation menu. Note that the default stop time is 10.0. Next, choose Run from the Simulation menu. Double-click the Scope block to see the sine wave signal as well as its integral (Figure 1.12). Obviously, the simulation stops when the stop time of 10.0 is reached. The simulation thus obtained cannot be copied and pasted into a document and is only for observation while in a MATLAB session. To gain access to the actual output data, the following needs be done. Select the Out1 block from Commonly Used Blocks library, and copy it onto the existing model. Then, draw the appropriate branch line to obtain the completed diagram shown in Figure 1.13. Once again, run the simulation. As a result, the time vector is automatically saved in tout, whereas the output is saved in yout. Next, at the MATLAB command prompt, type

```
>> plot(tout, yout)
```

This yields Figure 1.14. Note that this is the same result as observed in the simulation.

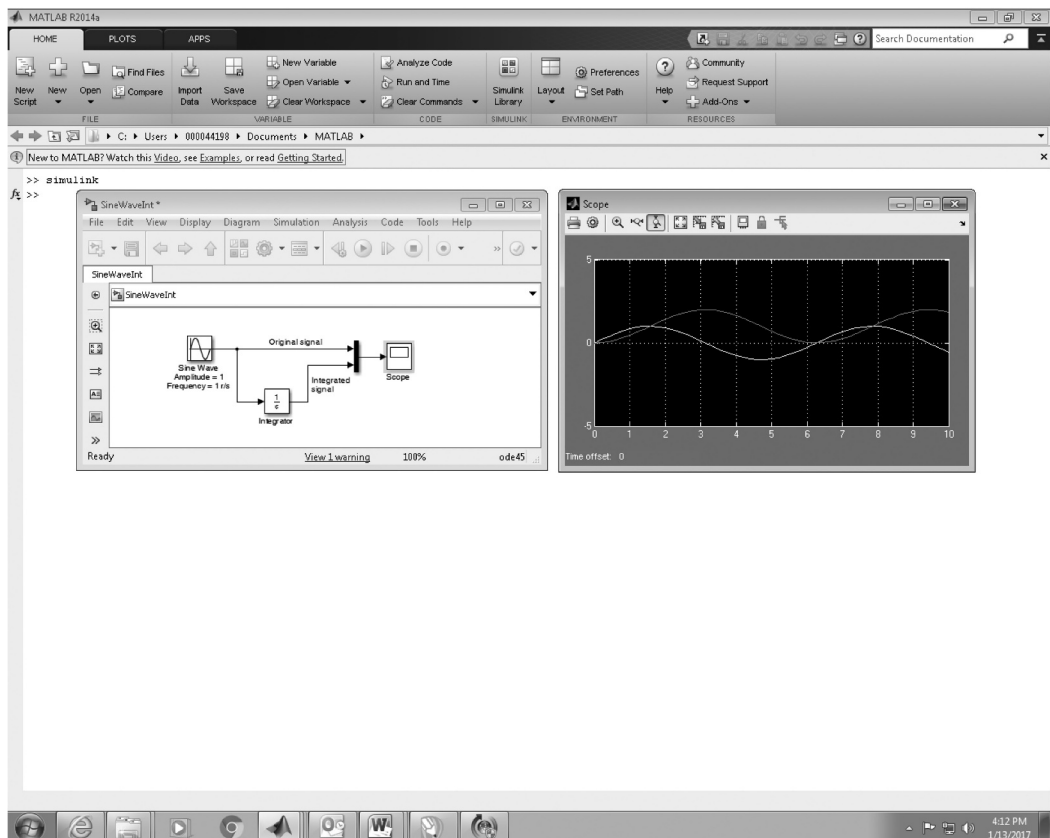


FIGURE 1.12
Simulation of a model.

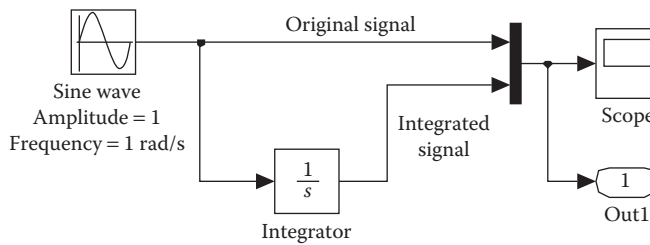


FIGURE 1.13
Storing output data.

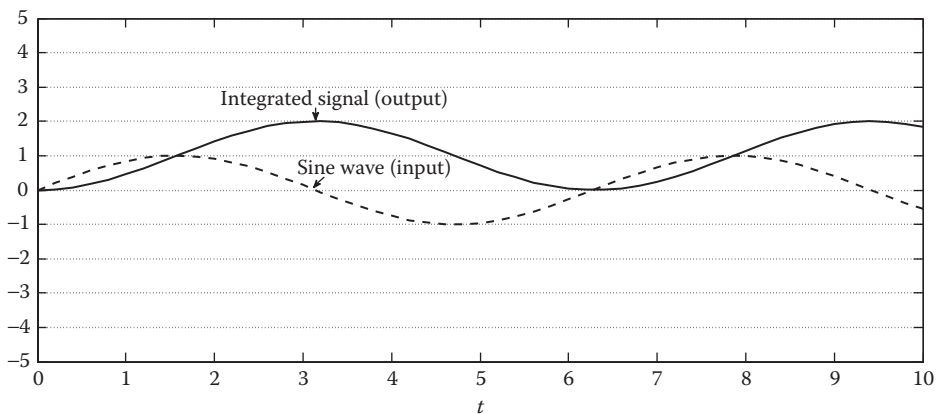


FIGURE 1.14
Output data.

1.7 Simscape

Simscape is a powerful software package that extends the Simulink product line with tools for modeling and simulation of physical systems, such as those with mechanical, electrical, hydraulic, thermal, and pneumatic components. Unlike other Simulink blocks, which represent mathematical operations or operate on signals, Simscape blocks directly represent physical components or relationships. With Simscape blocks, a system model is built in the same way as a physical system is assembled.

Simscape models use a Physical Network approach to model building: components (blocks) corresponding to physical elements, such as pumps, motors, and op-amps, are joined by lines corresponding to the physical connections that transmit power. This approach allows for the description of the physical structure of a system rather than the causal mathematics. Simscape automatically constructs, from the model, equations that characterize the system behavior, which are in turn integrated with the rest of the Simulink model. Simscape functions and utilities support functionality common to other Simulink products that use physical connections between their blocks. Simscape serves as the platform product for these add-on products of the Physical Modeling family:

- SimHydraulics®, for modeling and simulating hydraulic systems
- SimDriveline™, for modeling and simulating powertrain systems

- SimMechanics™, for modeling and simulating general mechanical systems
- SimElectronics®, for modeling and simulating electromechanical and electronic systems
- SimPowerSystems™, for modeling and simulating electrical power systems

These products can be used together to model physical systems in the Simulink environment.

1.7.1 Block Library

Type Simulink at the MATLAB command prompt to view the Simulink Library Browser, as previously shown in [Figure 1.5](#). Next, expand the Simscape entry in the contents tree ([Figure 1.15](#)). The Simscape block library contains two top-level libraries: Foundation Library and Utilities. If any add-on Physical Modeling products have been installed, they will appear under Simscape library. Each library can be expanded by double-clicking its icon. Double-clicking the Foundation Library icon and Utilities icon results in the menus displayed in [Figures 1.16](#) and [1.17](#), respectively.

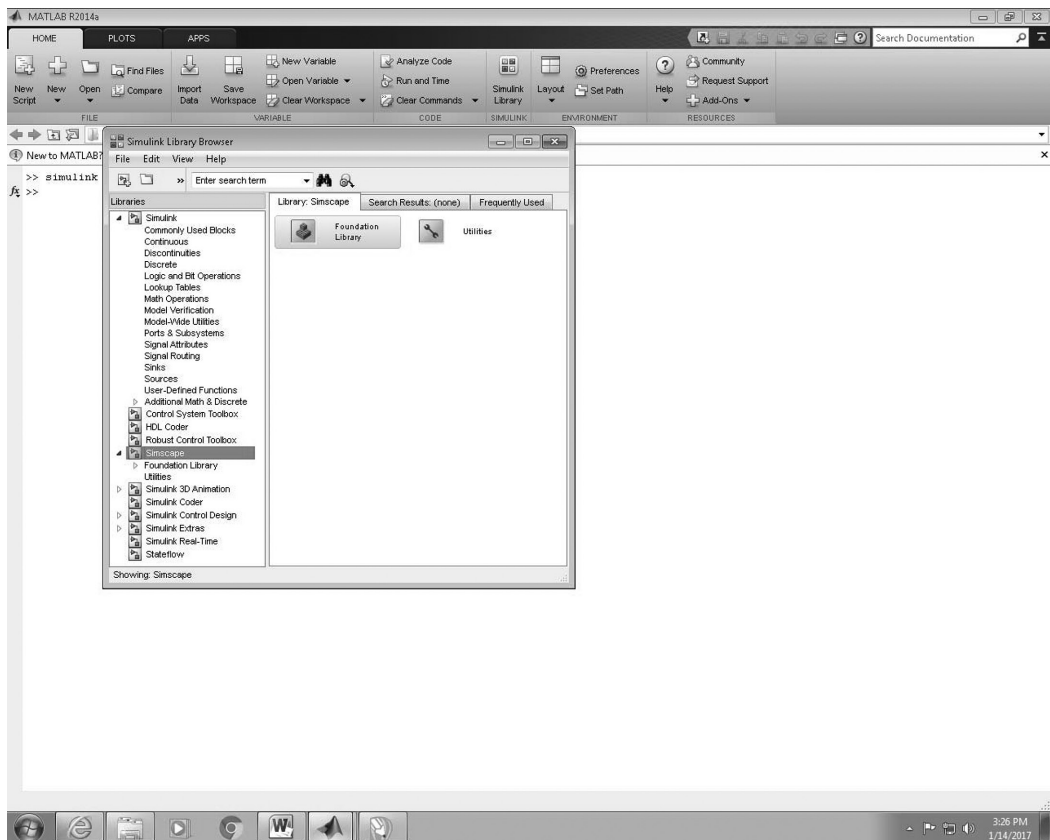


FIGURE 1.15
Simscape libraries.

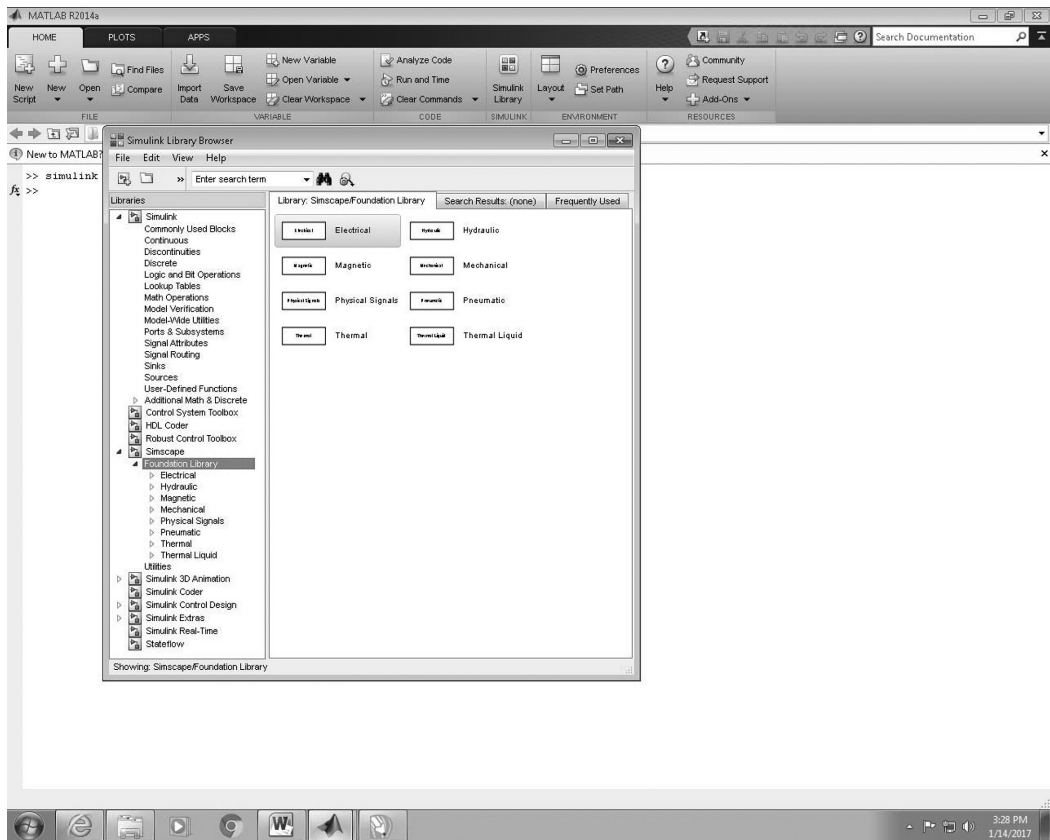


FIGURE 1.16
Sinscape Foundation library contents.

1.7.2 Building a New Model

A physical model can be built by using a combination of blocks from the Sinscape Foundation and Utilities libraries. Each Sinscape diagram must contain a Solver Configuration block from the Sinscape Utilities library (Figure 1.17). Regular Simulink blocks (such as sources or scopes) can be connected to the physical network diagram by using connector blocks: the Simulink-PS Converter block is used to connect Simulink outputs to Physical Signal inputs, and the PS-Simulink Converter block is used to connect Physical Signal outputs to Simulink inputs. Physical Signals, unlike Simulink signals, have units, which may be specified via the Sinscape block dialogs. Input and output signal units can be specified through the converter blocks.

Suppose we want to create a Sinscape diagram representing the RLC electrical circuit in Figure 1.18, which consists of a resistor, an inductor, and a capacitor and is driven by an applied voltage. We can vary the model parameters, such as the resistance or the applied voltage, and view the subsequent changes to the electric current. Suppose the input (applied voltage) is a unit pulse between $t = 1$ and $t = 2$, and the physical parameter values are $R = 1\ \Omega$, $L = 1\ \text{H}$, and $C = 0.2\ \text{F}$.

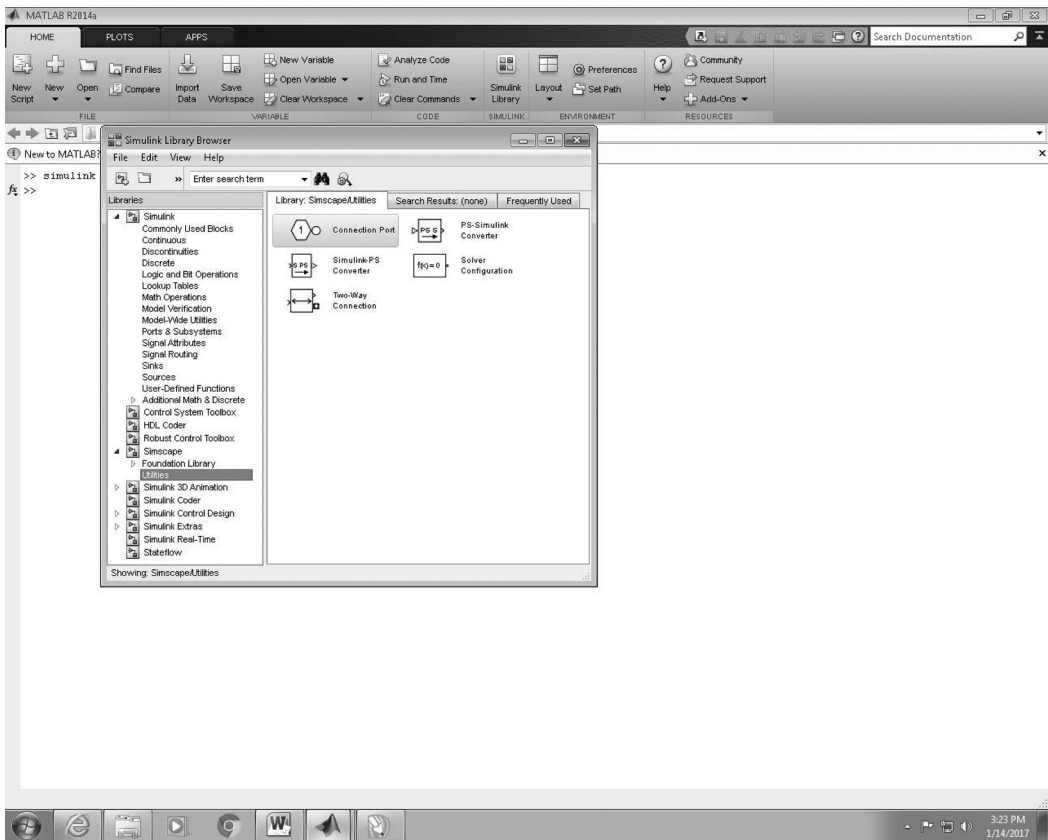


FIGURE 1.17
Simscape Utilities library contents.

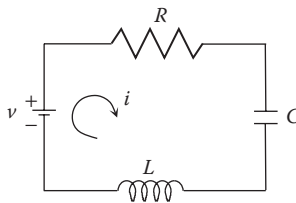


FIGURE 1.18
An RLC circuit.

The completed block diagram of this model will resemble [Figure 1.19](#). As before, to create a new model, select **File > New > Model** in the Simulink library window or simply press the **New Model** button (the icon underneath the **File** menu button) on the Library Browser's toolbar. This will open a new model editor window. Open the **Simscape > Foundation Library > Electrical > Electrical Elements** library. Drag the Capacitor,

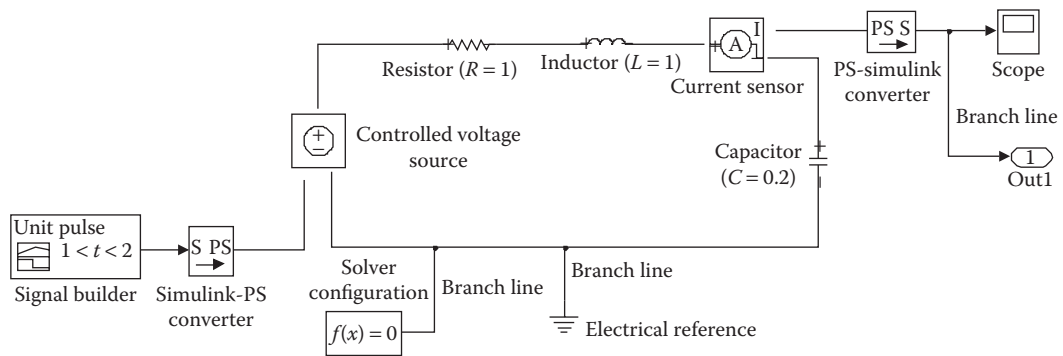


FIGURE 1.19
Simscape model for an RLC circuit.

Resistor, Inductor, and (one) Electrical Reference blocks into the model window. Note that the capacitor must be rotated 90 degrees clockwise to agree with what is shown in [Figure 1.19](#). This can be done by selecting the capacitor and choosing **Diagram > Rotate & Flip > Clockwise**. The representation of the applied voltage can be added by opening the **Simscape > Foundation Library > Electrical > Electrical Sources** library and adding the **Controlled Voltage Source** to the diagram. The current sensor can be added by opening the **Simscape > Foundation Library > Electrical > Electrical Sensors** library. Note that the *current sensor must be connected with the electrical elements in series*. The **Solver Configuration** block and the **PS-Simulink** and **Simulink-PS** converters are found in the **Simscape Utilities** library. For simulation purposes, a **Signal Builder** block from the **Sources** menu of **Simulink** is added to the model. This block will be used to define the input signal, which, in this case, is a unit pulse between $t = 1$ and $t = 2$. Double-click the **Signal Builder** block and then select, drag, and place the two vertical lines at $t = 1$ and $t = 2$. Make sure that the amplitude is 1 over this specified range. The completed model is shown in [Figure 1.19](#). Detailed information on modeling and analysis of electrical networks by using Simscape is presented in [Chapter 6](#).

1.7.3 Simulation

Select a Simulink solver. On the top menu bar of the model window, select **Simulation > Model Configuration Parameters**. The **Configuration Parameters** dialog box opens. Under **Solver** options, set **Solver** to **ode15s (Stiff/NDF)**, which is recommended for solving stiff differential equations, and set maximum step size to 0.2. Choose **Run** from the **Simulation** menu. Double-click the **Scope** block to see the output ([Figure 1.20](#)). As explained in [Section 1.6](#), the simulation thus obtained cannot be copied and pasted into a document and is only for observation while in a MATLAB session. To gain access to the actual output data, the **Out1** block from the **Commonly Used Blocks** library must be added to the existing model. Typing `plot(tout, yout)` at the MATLAB command prompt yields [Figure 1.21](#).

A wide range of dynamic systems can be modeled and analyzed using Simscape. As mentioned earlier, [Chapter 6](#) gives detailed coverage of Simscape modeling of electrical circuits. Simscape modeling of mechanical and thermal systems is presented in [Chapters 5](#) and [7](#), respectively.

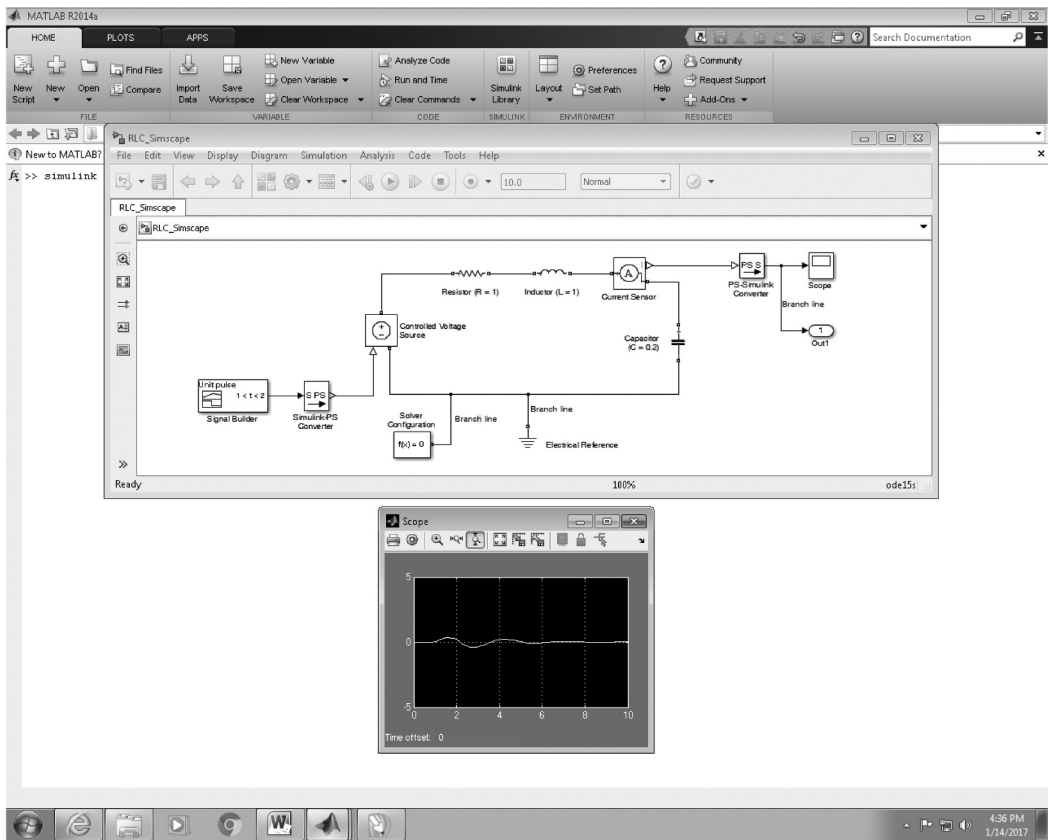


FIGURE 1.20
Simulation of the RLC circuit.

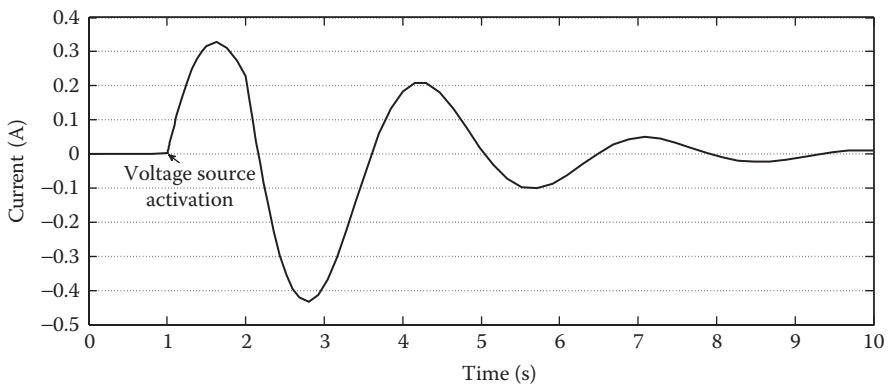


FIGURE 1.21
Output of the simulation of the RLC circuit.

Review Problems

- Evaluate the function $f(x, y) = \frac{2}{3} e^{-x/2} \cos(y - 1)$ for $x = -0.23, y = 2.7$ by:
 - Using the subs command.
 - Conversion into a MATLAB function.
- Evaluate the function $g(x, y) = \sin(2x + 1) \tan(y - \frac{1}{2})$ for $x = 0.45, y = -1.17$ by:
 - Using the subs command.
 - Conversion into a MATLAB function.
- Evaluate the vector function $v(x, y) = \begin{Bmatrix} xy + 1 \\ x - 2y \end{Bmatrix}$ for $x = 1.54, y = 2.28$ by:
 - Using the subs command.
 - Conversion into a MATLAB function.
- Evaluate the matrix function $m(x, y) = \begin{bmatrix} y & 3x - y \\ x + 1 & \sin 2y \end{bmatrix}$ for $x = -2, y = 3.35$ by:
 - Using the subs command.
 - Conversion into a MATLAB function.
- If $f(t) = e^{-3t/5} + t \ln(t + 1)$, evaluate df/dt when $t = 4.4$, by:
 - Using the subs command.
 - Conversion into a MATLAB function.
- If $g(x) = 2^{3x-1} + e^{-x} \cos x$, evaluate dg/dx when $x = 1.37$, by:
 - Using the subs command.
 - Conversion into a MATLAB function.
- Solve the following initial-value problem, and evaluate the solution at $x = 3.5$.

$$(x - 1)y' + xy = 2x, \quad y(2) = 1$$

- Solve the following initial-value problem, and evaluate the solution at $t = 2.8$.

$$t\dot{x} + x = e^t, \quad x(1) = -1$$

- Plot $y_1(t) = e^{-t/3} \cos(\frac{1}{2}t)$ and $y_2(t) = (t + 1)e^{-t}$ versus $0 \leq t \leq 10$ in the same graph. Adjust the limits on the vertical axis to -0.3 and 1.1 . Add grid and label.
- Plot $x_{1,2,3}(t) = e^{-at/2} \sin(\frac{1}{3}t)$, corresponding to $a = 1, 2, 3$, versus $0 \leq t \leq 5$ in the same graph. Adjust the limits on the vertical axis to -0.05 and 0.3 . Add grid and label.
- Plot $\int_1^t e^{x-t} \sin x \, dx$ versus $-1 \leq t \leq 3$. Add grid and label.
- Plot $\int_0^t (x + t)^2 e^{-(2t-x)} \, dx$ versus $1 \leq t \leq 3$. Add grid and label.
- Plot $z(x, t) = (1 + \cos x)e^{-(t-1)}$ versus $0 \leq x \leq 10$ for two values of $t = 1.5, 2.5$ in a 1×2 tile. Add grid and title.
- Plot $z(x, t) = \sin(2.7x) \cos(3.2t)$ versus $0 \leq x \leq 5$ for four values of $t = 1, 1.5, 2, 2.5$ in a 2×2 tile. Add grid and title.
- Given $f(x) = e^{-3x/2} + \cos(x - 1)$, plot $f'(x)$ versus $0 \leq x \leq 8$. Add grid and title.

16. Write a user-defined function with function call `F=temp_conv(C)` that converts the temperature from Celsius C to Fahrenheit F . Execute the function for the case of $C = 27$.
17. Write a user-defined function with function call `s=speed_calc(x,t0)` that calculates the speed of a moving object, whose position is described by a symbolic function x , at a specified time t_0 . Execute the function to find the speed (at $t = 2.5$) of an object whose position is described by $x(t) = 0.75t^3 - 2t^2 + 1.4$.
18. Write a user-defined function with function call `fval = f_eval(f,a,b)`, where f is an anonymous function and a and b are constants such that $a < b$. The function calculates the midpoint m of the interval $[a,b]$ and returns the value of $\frac{1}{2}f(a) + f(m) + \frac{1}{2}f(b)$. Execute `f_eval` for $f = xe^{-x}$, $a = -3$, $b = 4$.
19. Repeat the example considered in this chapter involving a signal and its integral, but this time, the sine wave has amplitude 0.8 and frequency 1.5 rad/sec and the integrated signal gets amplified by a factor of 2; see Figure 1.22. Build the model, run the simulation, and generate a plot that can be imported into a document.
20. Create the Simulink model shown in Figure 1.23. Select the signal generator as a square wave with amplitude 0.5 and frequency 0.5 rad/sec. To flip the gain block, right click on it, then go to Rotate & Flip, and choose the “Flip Block” option. Double-clicking the Sum (Commonly Used Blocks) allows control over the list of desired signs. Run simulation and generate a figure that can be imported into a document.

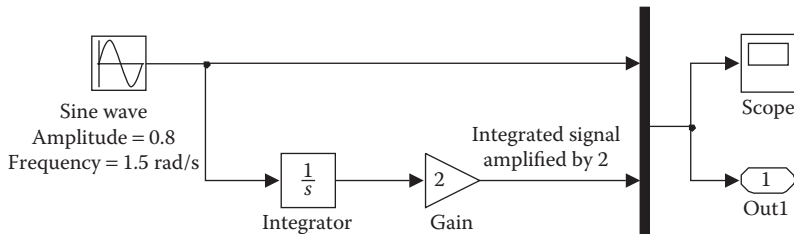


FIGURE 1.22
Problem 19.

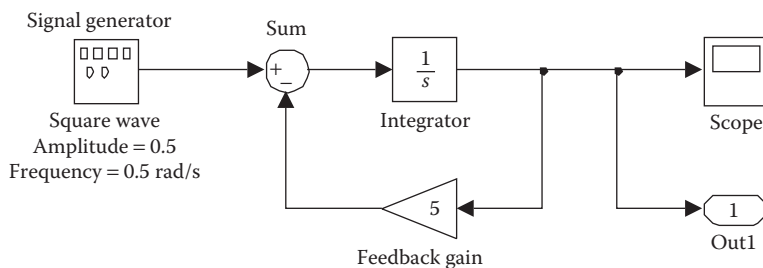


FIGURE 1.23
Problem 20.

21. Build the Simulink model shown in [Figure 1.24](#). Select the signal generator as a square wave with amplitude 1 and frequency 0.5 rad/sec. To flip the gain block, right click on it, then go to Rotate & Flip, and choose the “Flip Block” option. Double-clicking the Sum (Commonly Used Blocks) allows control over the list of desired signs. Run simulation and generate a figure that can be imported into a document.
22. Build the Simulink model shown in [Figure 1.25](#). To flip the gain block, right click on it, then go to Rotate & Flip, and choose the “Flip Block” option. Double-clicking the Sum allows control over the list of desired signs. Run simulation and generate the response plot.
23. Build the Simulink model shown in [Figure 1.26](#). Transfer function represents the ratio of the Laplace transforms of the output and the input signals associated with a block; see [Chapter 4](#). The Transfer Fcn block is located in the Continuous library. Run simulation and generate the response plot.

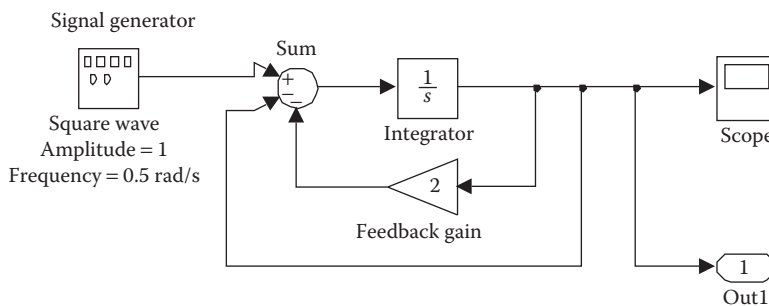


FIGURE 1.24
Problem 21.

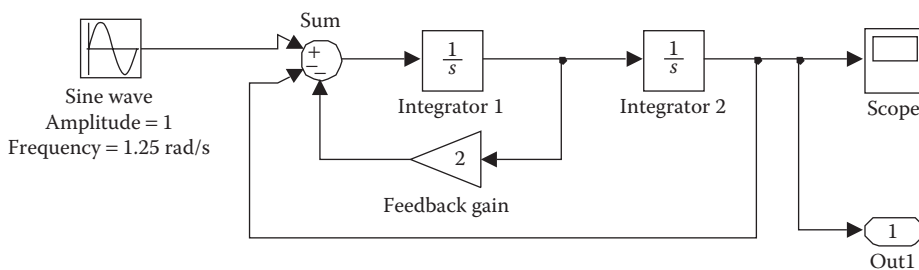


FIGURE 1.25
Problem 22.

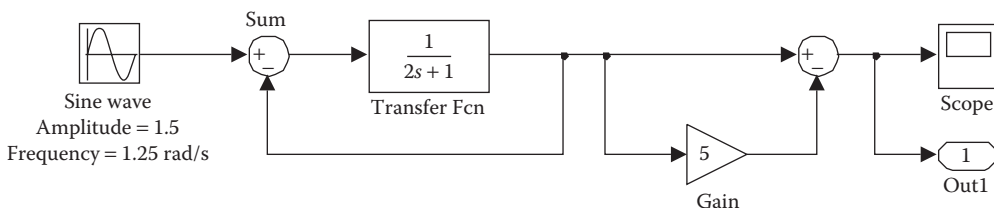


FIGURE 1.26
Problem 23.

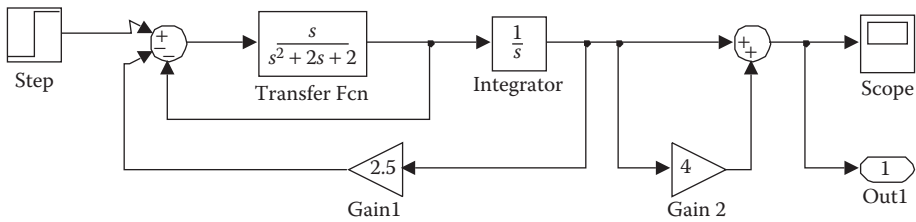


FIGURE 1.27
Problem 24.

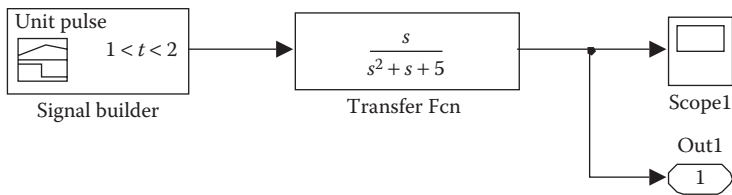


FIGURE 1.28
Problem 26.

24. Build the Simulink model shown in [Figure 1.27](#). The Step signal has a step time of 1 and a final value of 1; that is, it assumes a value of 1 immediately after $t = 1$ and a value of 0 otherwise. The Transfer Fcn block is located in the Continuous library. Run simulation and generate the response plot.
25. Build and simulate a Simulink model that integrates a sawtooth signal (amplitude 1 and frequency 1 rad/sec) and displays the result, along with the sawtooth signal itself.
26. Build the model shown in [Figure 1.28](#), which is the Simulink model of the RLC circuit considered in this chapter and is equivalent to the Simscape model presented in [Figure 1.19](#). Perform the simulation to confirm that both models yield the same response.
27. Build a Simscape model for an RL circuit ($R = 5 \, \Omega$, $L = 0.5 \, \text{H}$), driven by an applied voltage represented by a pulse (amplitude 0.6 for $1 < t < 3$). Run the simulation and generate the response (electric current) plot.
28. Build a Simscape model for an RC circuit ($R = 2 \, \Omega$, $C = 0.35 \, \text{F}$), driven by an applied voltage represented by a pulse (amplitude 1.4 for $1.5 < t < 2.5$). Run the simulation and generate the response (electric current) plot.



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2

Complex Analysis, Differential Equations, and Laplace Transformation

This chapter presents a review of complex analysis, differential equations, and Laplace transformation, providing the necessary background for a better understanding of various ideas and implementation of methods involved in the analysis of dynamic systems. Complex analysis comprises the study of complex numbers, complex variables, and complex functions. Ordinary differential equations (ODEs) arise in situations where the rate of change of a function with respect to its independent variable is involved. Differential equations are generally very difficult to solve, even for the simplest case of constant coefficients. To that end, Laplace transformation is used to solve initial-value problems (IVPs)—ODEs subjected to initial conditions—by transforming the data from time domain to the s -domain, where equations are algebraic and hence easier to work with. Transformation of the information from the s -domain back to time domain ultimately describes the solution of the IVP.

2.1 Complex Analysis

Complex analysis consists of the study of complex numbers, complex variables, and complex functions and plays an important role in a wide range of areas in the analysis of dynamic systems, from the calculation of a system's natural frequencies to the analysis of a system's frequency response.

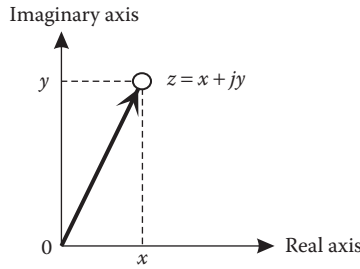
2.1.1 Complex Numbers in Rectangular Form

A complex number z in rectangular (Cartesian) form is expressed as

$$z = x + jy, \quad j = \sqrt{-1} = \text{imaginary number} \quad (2.1)$$

where x and y are real numbers, known as the real and imaginary parts of z , respectively, and are denoted by $x = \text{Re}\{z\}$ and $y = \text{Im}\{z\}$. For example, if $z = 2 - 3j$, then $x = \text{Re}\{z\} = 2$ and $y = \text{Im}\{z\} = -3$. A complex number with zero real part is called pure imaginary, for example, $z = 2j$. Two complex numbers are equal if and only if their respective real and imaginary parts are equal. Addition of complex numbers is performed component-wise, that is, if $z_1 = x_1 + jy_1$ and $z_2 = x_2 + jy_2$, then

$$\begin{aligned} z_1 + z_2 &= (x_1 + jy_1) + (x_2 + jy_2) \\ &= (x_1 + x_2) + j(y_1 + y_2) \end{aligned}$$

**FIGURE 2.1**

Geometry of a complex number.

Multiplication of two complex numbers is performed in the same way as two binomials, with the provision that $j^2 = -1$, that is,

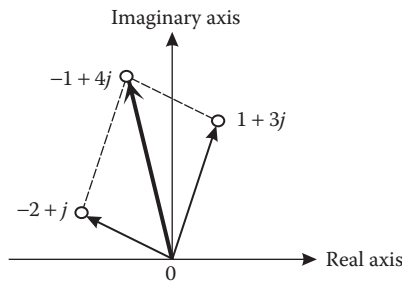
$$\begin{aligned} z_1 z_2 &= (x_1 + jy_1)(x_2 + jy_2) = x_1 x_2 + jy_1 x_2 + jx_1 y_2 + j^2 y_1 y_2 \\ &= (x_1 x_2 - y_1 y_2) + j(x_1 y_2 + x_2 y_1) \end{aligned}$$

Complex numbers have a two-dimensional character, because they consist of a real part and an imaginary part. Therefore, they may be represented geometrically as points in a Cartesian coordinate system, known as the complex plane. The x -axis of the complex plane is the real axis, and the y -axis is the imaginary axis (Figure 2.1). Since $z = x + jy$ is uniquely identified by an ordered pair (x, y) of real numbers, it can be represented as a position vector in the complex plane, with initial point 0 and terminal point $z = x + jy$. The concept of vector addition therefore applies to the addition of complex numbers. For example, if $z_1 = 1 + 3j$ and $z_2 = -2 + j$, as in Figure 2.2, then $z_1 + z_2 = -1 + 4j$ agrees with the resultant of the position vectors of z_1 and z_2 .

2.1.1.1 Magnitude

The magnitude (or modulus) of a complex number $z = x + jy$ is denoted by $|z|$ and is defined as

$$|z| = \sqrt{x^2 + y^2} \quad (2.2)$$

**FIGURE 2.2**

Addition of complex numbers as vectors.

Referring to [Figure 2.1](#), the magnitude of z is simply the distance from the origin to z . If z is a real number ($z = x$), it is located on the real axis and its magnitude is equal to its absolute value. If z is pure imaginary ($z = jy$), it is on the imaginary axis and $|z| = |y|$. The distance between two complex numbers z_1 and z_2 is given by $|z_1 - z_2|$. Addition of complex numbers follows the triangle inequality rule,

$$|z_1 + z_2| \leq |z_1| + |z_2| \quad (2.3)$$

Example 2.1: Magnitude

Let $z_1 = -2 + 3j$, $z_2 = 1 + 4j$.

- Find the distance between z_1 and z_2 .
- Verify the triangle inequality.
-  Confirm the results of (a) and (b) in MATLAB®.

Solution

- Noting that $z_1 - z_2 = -3 - j$, we find $|z_1 - z_2| = \sqrt{(-3)^2 + (-1)^2} = \sqrt{10}$.
- To verify the triangle inequality, Equation 2.3, we first calculate

$$|z_1| = |-2 + 3j| = \sqrt{13}, \quad |z_2| = |1 + 4j| = \sqrt{17}, \quad |z_1 + z_2| = |-1 + 7j| = \sqrt{50}$$

It is then readily seen that $\sqrt{50} \leq \sqrt{13} + \sqrt{17}$.

- 

```
% MATLAB recognizes both i and j as the imaginary number
>> z1 = -2+3*j; z2 = 1+4*j;
>> abs(z1 - z2)
ans =
    3.1623          % sqrt(10)

>> abs(z1 + z2) - abs(z1) - abs(z2)
ans =
   -0.6576          % Negative result confirms triangle inequality
```

2.1.1.2 Complex Conjugate

The complex conjugate of $z = x + jy$ is denoted by $\bar{z}$ and defined as $\bar{z} = x - jy$. The product of a complex number ($z \neq 0$) and its conjugate is a positive real number, equal to the square of the magnitude of z ,

$$z\bar{z} = (x + jy)(x - jy) = x^2 + y^2 = |z|^2 \quad (2.4)$$

Geometrically, $\bar{z}$ is the reflection of z through the real axis. Conjugates play an important role in the division of complex numbers. Let us consider z_1/z_2 , where $z_1 = x_1 + jy_1$ and

$z_2 = x_2 + jy_2$ ($z_2 \neq 0$). Multiplication of the numerator and the denominator by the conjugate of the denominator ($\bar{z}_2 = x_2 - jy_2$) results in

$$\begin{aligned} \frac{x_1 + jy_1}{x_2 + jy_2} &= \frac{(x_1 + jy_1)(x_2 - jy_2)}{(x_2 + jy_2)(x_2 - jy_2)} \stackrel{\text{Using Equation 2.4}}{=} \frac{(x_1x_2 + y_1y_2) + j(y_1x_2 - y_2x_1)}{x_2^2 + y_2^2} \\ &= \frac{x_1x_2 + y_1y_2}{x_2^2 + y_2^2} + j \frac{y_1x_2 - y_2x_1}{x_2^2 + y_2^2} \end{aligned}$$

Note that the result has been expressed in standard rectangular form.

Example 2.2: Conjugation

- a. Express the following in rectangular form:

$$\frac{\frac{1}{2} - j}{-1 + \frac{1}{3}j}$$

- b.  Repeat (a) in MATLAB.

Solution

a.
$$\frac{\frac{1}{2} - j}{-1 + \frac{1}{3}j} = \frac{(\frac{1}{2} - j)(-1 - \frac{1}{3}j)}{(-1 + \frac{1}{3}j)(-1 - \frac{1}{3}j)} = \frac{-\frac{5}{6} + \frac{5}{6}j}{\frac{10}{9}} = \frac{-3}{4} + \frac{3}{4}j$$

- b. 

Following the exact steps as in the solution earlier, and using the `conj` command, we find

```
>> z1 = 1/2-j; z2 = -1+1/3*j; z1*conj(z2)/(z2*conj(z2))
ans =
-0.7500 + 0.7500i    % Note that MATLAB returns i instead of j
```

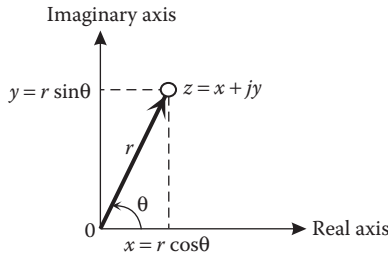
Direct calculation yields the same result:

```
>> z1/z2
ans =
-0.7500 + 0.7500i
```

2.1.2 Complex Numbers in Polar Form

The standard rectangular form $z = x + jy$ is quite inefficient in many situations, for instance, in the calculation of $(-1 + 3j)^8$. To remedy this, the polar form of a complex number is utilized. As its name suggests, the polar form uses polar coordinates to represent a complex number in the complex plane. The location of any point $z = x + jy$ in the complex plane can be determined by a radial coordinate r and an angular coordinate θ . The relationships between the rectangular and polar coordinates (Figure 2.3) are described by

$$x = r \cos \theta, \quad y = r \sin \theta$$

**FIGURE 2.3**

Relation between rectangular and polar coordinates.

We first introduce Euler's formula,

$$e^{j\theta} = \cos \theta + j \sin \theta \quad (2.5)$$

Consequently,

$$z = x + jy = \underbrace{r \cos \theta}_x + j \underbrace{r \sin \theta}_y = r (\cos \theta + j \sin \theta) = r e^{j\theta} \quad (2.6)$$

The result, $z = r e^{j\theta}$, is known as the polar form of z . Here, r is the magnitude of z , given by

$$r = |z| = \sqrt{x^2 + y^2}$$

and the phase (argument) of z is determined by

$$\theta = \arg z = \tan^{-1} \left(\frac{\text{Im}\{z\}}{\text{Re}\{z\}} \right) = \tan^{-1} \left(\frac{y}{x} \right) \quad (2.7)$$

The angle θ is measured from the positive real axis and, by convention, is regarded as positive in the counterclockwise direction. It is measured in radians (rad) and is determined in terms of integer multiples of 2π . The specific value of θ that lies in the interval $(-\pi, \pi]$ is called the principal value of $\arg z$ and is denoted by $\text{Arg } z$. In engineering applications, it is also common to express the polar form as

$$z = r \angle \theta$$

where $\angle$ denotes the phase angle. Note that the phase angle is not fully determined by Equation 2.7. As the following example demonstrates, the complex number needs to be located before its phase may be assessed.

Example 2.3: Polar Form

Express $z = -1 - 2j$ in polar form.

Solution

Location of z , [Figure 2.4](#), will facilitate the phase calculation. By Equation 2.7,

$$\theta = \tan^{-1}\left(\frac{-2}{-1}\right) \cong 1.1071 \text{ rad} \cong 63.4349^\circ$$

However, this is not the phase. Because z is in the third quadrant, the phase in the positive sense (counterclockwise) is calculated as $1.1071 + \pi \cong 4.2487$ rad, whereas in the negative sense (clockwise), it is $\pi - 1.1071 \cong 2.0344$ rad, which has the smaller measure between the two. Noting that $r = \sqrt{5}$, and using the phase in the negative direction, we find

$$z = -1 - 2j = \sqrt{5} e^{-2.0344j}$$

2.1.2.1 Complex Algebra Using the Polar Form

Working with the polar form considerably simplifies complex algebra. Consider two complex numbers, in their respective polar forms, $z_1 = r_1 e^{j\theta_1}$ and $z_2 = r_2 e^{j\theta_2}$. Then,

$$z_1 z_2 = r_1 r_2 e^{j(\theta_1 + \theta_2)} \quad \text{or alternatively,} \quad r_1 r_2 \angle (\theta_1 + \theta_2)$$

This means that the magnitude and phase of the product $z_1 z_2$ are

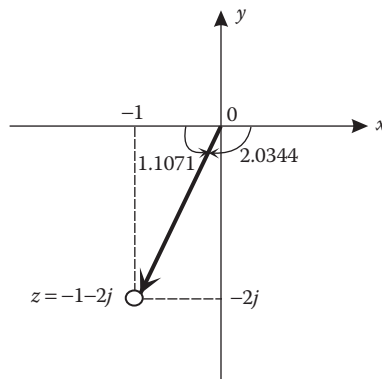
$$|z_1 z_2| = r_1 r_2 = |z_1| |z_2| \quad \text{and} \quad \arg(z_1 z_2) = \theta_1 + \theta_2 = \arg(z_1) + \arg(z_2)$$

Similarly, in the case of division,

$$\frac{z_1}{z_2} = \frac{r_1}{r_2} e^{j(\theta_1 - \theta_2)} \quad \text{or} \quad \frac{r_1}{r_2} \angle (\theta_1 - \theta_2)$$

Therefore, the magnitude and phase of z_1 / z_2 are

$$\left| \frac{z_1}{z_2} \right| = \frac{r_1}{r_2} = \frac{|z_1|}{|z_2|} \quad \text{and} \quad \arg\left(\frac{z_1}{z_2}\right) = \theta_1 - \theta_2 = \arg(z_1) - \arg(z_2)$$

**FIGURE 2.4**

Calculation of phase in Example 2.3.

Example 2.4: Division Using Polar Form

Express the following in polar form:

$$\frac{-1-2j}{-1+j}$$

Solution

The polar form of the numerator was derived in Example 2.3, as

$$-1-2j = \sqrt{5} e^{-2.0344j}$$

The denominator is in the second quadrant and its polar form is

$$-1+j = \sqrt{2} e^{(3\pi/4)j}$$

As a result,

$$\frac{-1-2j}{-1+j} = \frac{\sqrt{5} e^{-2.0344j}}{\sqrt{2} e^{(3\pi/4)j}} = \sqrt{\frac{5}{2}} e^{-4.3906j}$$

This may be verified as follows:

$$\frac{-1-2j}{-1+j} \cdot \frac{-1-j}{-1-j} = \frac{-1+3j}{2} = \frac{-1}{2} + \frac{3}{2}j \quad \begin{array}{l} \text{Polar form} \\ \text{second quadrant} \end{array} = \sqrt{\frac{5}{2}} e^{1.8925j} \quad \begin{array}{l} 2\pi - 1.8925 = -4.3906 \\ = \end{array} \sqrt{\frac{5}{2}} e^{-4.3906j}$$

Given $z = re^{j\theta}$, its conjugate is derived as

$$\begin{aligned} \bar{z} &= x - jy = r \cos \theta - j(r \sin \theta) = r(\cos \theta - j \sin \theta) = r[\cos(-\theta) + j \sin(-\theta)] \\ &\stackrel{\text{Euler's formula}}{=} re^{-j\theta} \end{aligned}$$

This makes sense geometrically, since a complex number and its conjugate are reflections of one another through the real axis. Hence, they are equidistant from the origin, that is, $|z| = |\bar{z}| = r$, and their phases are equal in measure but opposite in sign, that is, $\arg(z) = -\arg(\bar{z})$ (Figure 2.5). The magnitude property of complex conjugation, Equation 2.4, can also be confirmed, as

$$z\bar{z} = (re^{j\theta})(re^{-j\theta}) = r^2 = |z|^2$$

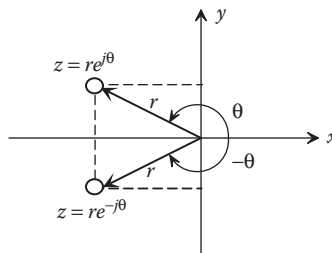


FIGURE 2.5

A complex number and its conjugate.

2.1.2.2 Integer Powers of Complex Numbers

As mentioned earlier, one area that reveals the effectiveness of the polar form is in dealing with expressions in the form z^n for integer n . The idea is as follows:

$$\begin{aligned}
 z^n & \stackrel{\text{Polar form}}{=} (re^{j\theta})^n = r^n e^{jn\theta} \stackrel{\text{Euler's formula}}{=} r^n (\cos n\theta + j \sin n\theta) \\
 & \stackrel{\text{Rectangular form}}{=} r^n \cos n\theta + jr^n \sin n\theta
 \end{aligned} \tag{2.8}$$

Example 2.5: Integer Power

Simplify $(-1 - 2j)^4$.

Solution

By Example 2.3, we have $-1 - 2j = \sqrt{5} e^{-2.0344j}$. Using this in Equation 2.8, with $n = 4$,

$$(-1 - 2j)^4 = \left(\sqrt{5} e^{-2.0344j} \right)^4 = 25 e^{-8.1376j} = -7 - 24j$$

2.1.2.3 Roots of Complex Numbers

In real calculus, if a is a real number, then $\sqrt[n]{a}$ has a single value. On the contrary, given a complex number $z \neq 0$ and a positive integer n , the n th root of z , written $\sqrt[n]{z}$, is *multi-valued*. In fact, there are n different values of $\sqrt[n]{z}$ corresponding to each value of $z \neq 0$. If $z = re^{j\theta}$, it can then readily be shown that

$$\sqrt[n]{z} = \sqrt[n]{r} \left(\cos \frac{\theta + 2k\pi}{n} + j \sin \frac{\theta + 2k\pi}{n} \right), \quad k = 0, 1, \dots, n-1 \tag{2.9}$$

Geometrically, these n values lie on a circle centered at the origin, with a radius of $\sqrt[n]{r}$, and are the n vertices of an n -sided regular polygon.

Example 2.6: Third Roots of 1

Calculate all three values of $\sqrt[3]{1}$.

Solution

The objective is to find $w = \sqrt[3]{z}$, where $z = 1$. Noting that $z = 1$ is located on the positive real axis, one unit from the origin, we have $r = 1$ and $\theta = 0$; hence, $z = 1 = e^{j(0)}$. Using Equation 2.9 with $n = 3$, $r = 1$, and $\theta = 0$, we find

$$\sqrt[3]{1} = \sqrt[3]{1} \left(\cos \frac{0 + 2k\pi}{3} + j \sin \frac{0 + 2k\pi}{3} \right), \quad k = 0, 1, 2$$

This generates three complex numbers, each with a magnitude of 1. The phases associated with $k = 0, 1, 2$ are $0, \frac{2}{3}\pi, \frac{4}{3}\pi$. Therefore, the three roots are found as $1, -\frac{1}{2} + \frac{\sqrt{3}}{2}j, -\frac{1}{2} - \frac{\sqrt{3}}{2}j$. Note that all three roots lie on a circle of radius $\sqrt[3]{1} = 1$ centered at the origin and are the vertices of a regular three-sided polygon, as asserted.

2.1.3 Complex Variables and Complex Functions

If x or y or both vary, then $z = x + iy$ is called a complex variable. The Laplace variable (Section 2.3) is a recognized example of a complex variable. A complex function F defined on a set S is a rule that assigns a complex number w to each $z \in S$. The notation is $w = F(z)$, and the set S is the domain of definition of F . For instance, the domain of the function $F(z) = z/(z+1)$ is any region that excludes the point $z = -1$. Because z assumes different values from the set S , it is clearly a complex variable. Transfer functions (Section 4.3.2) and frequency response functions (Section 8.4) are the examples of complex functions that frequently arise in the analysis of dynamic systems.

PROBLEM SET 2.1

In Problems 1 through 4,

a. Express z_1/z_2 in rectangular form.

b. Verify that $|z_1/z_2| = |z_1|/|z_2|$.

1. $\frac{-2j}{3-4j}$

2. $\frac{1-j}{2-\frac{1}{2}j}$

3. $\frac{\frac{1}{3}}{-2+j}$

4. $\frac{\frac{2}{3}+2j}{-3j}$

In Problems 5 through 10, express each complex number in its polar form.

5. $\sqrt{3}-2j$

6. $-\frac{2}{3}+j$

7. $\sqrt{3}+j$

8. $-\frac{1}{2}j$

9. $\frac{1+j}{-1+2j}$

10. $\frac{1+j\sqrt{3}}{1-j\sqrt{3}}$

In Problems 11 through 15, perform the operations by using the polar form and express the result in rectangular form.

11. $\frac{\frac{2}{3}j}{1+j\sqrt{3}}$

12. $\frac{1-\frac{1}{3}j}{\frac{1}{3}+j}$

13. $\frac{4+j}{(4+3j)^3}$

14. $(0.9239 + 0.3827j)^{12}$

15. $\left(\frac{j}{1+4j}\right)^3$

In Problems 16 through 20, find all possible values for each expression.

16. $(-1+2j)^{1/3}$

17. $(-1)^{1/4}$

18. $(\sqrt{3}-j)^{1/4}$

19. $\sqrt{1+j\sqrt{2}}$

20. $(-\frac{2}{3}+j)^{2/3}$

21. Find all possible roots of $jz^3 - (1+j) = 0$.

2.2 Differential Equations

An equation involving an unknown function and one or more of its derivatives is called a differential equation. Differential equations are divided into two categories: ordinary differential equations (ODEs) and partial differential equations (PDEs). When there is only one independent variable, the equation is an ODE. For example, $\frac{1}{4}\dot{x} + x = e^{-t/2}$ is an ODE involving the unknown function $x(t)$, its first derivative $\dot{x} = dx/dt$, as well as a given function $e^{-t/2}$. Similarly, $2\ddot{x} + t\dot{x}^2 = \sin t$ is an ODE relating $x(t)$ and its first and second derivatives with respect to t , as well as the function $\sin t$. In dynamic system models, the independent variable is generally time t . The derivative of the highest order of the unknown function $x(t)$ with respect to t is the order of the ODE. For instance, $\frac{1}{4}\dot{x} + x = e^{-t/2}$ is a first-order differential equation and $2\ddot{x} + t\dot{x}^2 = \sin t$ is a second-order differential equation. Consider an n th-order ODE in the form

$$a_n x^{(n)} + a_{n-1} x^{(n-1)} + \dots + a_1 \dot{x} + a_0 x = F(t) \quad (2.10)$$

where $x = x(t)$ and $x^{(n)} = d^n x / dt^n$. If all coefficients $a_0, a_1, \dots, a_n$ are either constants or functions of the independent variable t , the ODE is linear. Otherwise, the ODE is nonlinear. If $F(t) \equiv 0$, the ODE is homogeneous. Otherwise, it is nonhomogeneous. Therefore, $\frac{1}{4}\dot{x} + x = e^{-t/2}$ is linear, whereas $2\ddot{x} + t\dot{x}^2 = \sin t$ is nonlinear, and both are nonhomogeneous. If the dependent variable is a function of more than one independent variable, the equation is called a PDE.

2.2.1 Linear, First-Order Differential Equations

In agreement with Equation 2.10, a linear, first-order ODE is expressed as

$$a_1 \dot{x} + a_0 x = F(t) \quad \begin{array}{c} \text{Divide by } a_1 \\ \Rightarrow \\ \text{and rewrite as} \end{array} \quad \dot{x} + g(t)x = f(t) \quad (2.11)$$

with a general solution in the form

$$x(t) = e^{-h} \left[\int e^h f(t) dt + c \right], \quad h = \int g(t) dt \quad (2.12)$$

where c is a constant of integration. A particular solution can be obtained when an initial condition is specified. Assuming that the initial time is $t = t_0$, the initial condition for x refers to the value of x immediately prior to the initial time and is denoted by $x(t_0^-)$. On the other hand, the initial value of x , denoted by $x(t_0^+)$, is its value immediately after the initial time. Although the initial condition and initial value of a quantity are almost always the same, there are rare instances in dynamic systems analysis where they are different; see [Section 2.3](#). In the meantime, we simply assume that the two are the same and denoted by $x(t_0)$. That said, a first-order IVP is described by

$$\dot{x} + g(t)x = f(t), \quad x(t_0) = x_0$$

Example 2.7: Initial-Value Problem

- a. Solve the IVP

$$\frac{1}{3} \dot{x} + x = e^{-t/2}, \quad x(0) = \frac{1}{5}$$

- b.  Confirm the result of (a) in MATLAB.

Solution

- a. Rewrite the ODE as $\dot{x} + 3x = 3e^{-t/2}$, so that $g(t) = 3$ and $f(t) = 3e^{-t/2}$ in Equation 2.11. Then, by Equation 2.12, we have $h = \int 3dt = 3t$, and a general solution is

$$x(t) = e^{-3t} \left[\int 3e^{3t} e^{-t/2} dt + c \right] = e^{-3t} \left[\frac{6}{5} e^{5t/2} + c \right] = \frac{6}{5} e^{-t/2} + ce^{-3t}$$

Using the initial condition, we find $c = -1$, and consequently, $x(t) = \frac{6}{5} e^{-t/2} - e^{-3t}$.

- b. 

```
>> x = dsolve('(1/3)*Dx+x=exp(-t/2)', 'x(0)=1/5'); simplify(x)
ans =
(6*exp(-t/2))/5 - exp(-3*t)
```

2.2.2 Second-Order Differential Equations with Constant Coefficients

A special class of linear, second-order ODEs that arises in dynamic system models is in the form

$$\begin{array}{ccc} a_2 \ddot{x} + a_1 \dot{x} + a_0 x = F(t) & \begin{array}{c} \text{Divide by } a_2 \\ \Rightarrow \\ \text{and rewrite} \end{array} & \ddot{x} + b \dot{x} + cx = f(t) \end{array} \quad (2.13)$$

Constant coefficients

This equation is normally accompanied by a set of two initial conditions, $x(t_0)$ and $\dot{x}(t_0)$, where t_0 denotes the initial time. A general solution $x(t)$ of Equation 2.13 is a superposition of the homogeneous solution $x_h(t)$, sometimes called the complementary solution $x_c(t)$, and the particular solution $x_p(t)$, that is, $x(t) = x_h(t) + x_p(t)$.

2.2.2.1 Homogeneous Solution

The homogeneous solution $x_h(t)$ is the solution of the homogeneous equation,

$$\ddot{x} + b\dot{x} + cx = 0$$

and is found as follows. Assume that $x = e^{\lambda t}$, with λ to be determined, and this is inserted in the earlier equation:

$$(\lambda^2 + b\lambda + c)e^{\lambda t} = 0 \quad \xRightarrow{e^{\lambda t} \neq 0} \quad \lambda^2 + b\lambda + c = 0 \quad \xRightarrow{\text{Solve}} \quad \begin{matrix} \lambda_1 \\ \lambda_2 \end{matrix}$$

Characteristic equation

The characteristic values λ_1 and λ_2 establish the two linearly independent solutions that form $x_h(t)$.

Case (1): $\lambda_1 \neq \lambda_2$ (real)

The two independent solutions are $e^{\lambda_1 t}$ and $e^{\lambda_2 t}$, and a linear combination of the two yields

$$x_h(t) = c_1 e^{\lambda_1 t} + c_2 e^{\lambda_2 t}, \quad c_1, c_2 = \text{const}$$

Case (2): $\lambda_1 = \lambda_2 = \lambda$

The independent solutions are $e^{\lambda t}$ and $te^{\lambda t}$, so that

$$x_h(t) = (c_1 + c_2 t)e^{\lambda t}, \quad c_1, c_2 = \text{const}$$

Case (3) $\lambda_2 = \bar{\lambda}_1$ (complex conjugates)

If $\lambda_1 = \alpha + j\beta$, then the two independent solutions are $e^{\alpha t} \cos \beta t$ and $e^{\alpha t} \sin \beta t$, leading to

$$x_h(t) = e^{\alpha t} (c_1 \cos \beta t + c_2 \sin \beta t), \quad c_1, c_2 = \text{const}$$

2.2.2.2 Particular Solution

The particular solution $x_p(t)$ of Equation 2.13 is obtained by using the method of undetermined coefficients. The method is limited in its applications and only handles cases in which $f(t)$ is a polynomial, exponential, sinusoidal, or some combination of them. Table 2.1 contains the recommended $x_p(t)$ for different scenarios of $f(t)$. These recommended expressions are subject to modification in some special cases, as follows. If x_p

TABLE 2.1

Selection of Particular Solution; Method of Undetermined Coefficients

Term in $f(t)$	Recommended $x_p(t)$
$A_n t^n + \dots + A_1 t + A_0$	$K_n t^n + \dots + K_1 t + K_0$
Ae^{at}	Ke^{at}
$A \sin \omega t$ or $A \cos \omega t$	$K_1 \cos \omega t + K_2 \sin \omega t$
$Ae^{\sigma t} \sin \omega t$ or $Ae^{\sigma t} \cos \omega t$	$e^{\sigma t} (K_1 \cos \omega t + K_2 \sin \omega t)$

contains a term that coincides with an independent solution of the homogeneous equation and the solution corresponds to a nonrepeated characteristic value, then the recommended x_p must be multiplied by t . If the characteristic value is repeated—as in above-mentioned Case (2)—then x_p must be multiplied by t^2 .

Example 2.8: Nonhomogeneous, Second-Order Ordinary Differential Equation

a. Solve the IVP

$$3\ddot{x} + 5\dot{x} + 2x = e^{-2t/3}, \quad x(0) = 0, \quad \dot{x}(0) = 1$$

b.  Confirm the result of (a) in MATLAB.

Solution

a. The characteristic equation $3\lambda^2 + 5\lambda + 2 = 0$ yields $\lambda = -1, -\frac{2}{3}$, so that $x_h = c_1 e^{-t} + c_2 e^{-2t/3}$. Based on Table 2.1, we choose $x_p = K e^{-2t/3}$. However, $e^{-2t/3}$ coincides with the independent solution in x_h associated with $\lambda = -\frac{2}{3}$. Therefore, x_p is modified to $x_p = K t e^{-2t/3}$. Inserting into the original ODE and equating the two sides, we find $K = 1$, so that $x_p = t e^{-2t/3}$. This implies that a general solution is $x(t) = c_1 e^{-t} + c_2 e^{-2t/3} + t e^{-2t/3}$. Finally, application of the initial conditions results in $c_1 = c_2 = 0$; hence, $x(t) = t e^{-2t/3}$.

b. 

```
>> x = dsolve('3*D2x+5*Dx+2*x = exp(-2*t/3)', 'x(0)=0', 'Dx(0)=1');
>> simplify(x)
ans =
t*exp(-(2*t)/3)
```

Example 2.9: Homogeneous, Second-Order Ordinary Differential Equation

Solve the IVP

$$\ddot{x} + 2\dot{x} + 5x = 0, \quad x(0) = 1, \quad \dot{x}(0) = 3$$

Solution

The roots of the characteristic equation $\lambda^2 + 2\lambda + 5 = 0$ are $\lambda = -1 \pm 2j$ and a general solution is formed as $x(t) = e^{-t}(c_1 \cos 2t + c_2 \sin 2t)$. Applying the initial conditions leads to $c_1 = 1$ and $c_2 = 2$; hence, $x(t) = e^{-t}(\cos 2t + 2 \sin 2t)$.

2.2.2.2.1 Expressing $A \cos \omega t + B \sin \omega t$ as $D \sin(\omega t + \phi)$

In systems analysis, we often encounter expressions in the form $A \cos \omega t + B \sin \omega t$, where the sine and cosine waves have the same frequency ω , for example, the term $\cos 2t + 2 \sin 2t$ that appears in the solution of Example 2.9. However, it is more convenient to write this expression as a single trigonometric term such as $D \sin(\omega t + \phi)$, where D is the amplitude, ϕ is the phase shift, and ω matches the common frequency in the sine and cosine terms. Using a trigonometric expansion, we have

$$D \sin(\omega t + \phi) = D \boxed{\cos \omega t} \sin \phi + D \boxed{\sin \omega t} \cos \phi \quad \overset{\text{Compare}}{=} \quad A \boxed{\cos \omega t} + B \boxed{\sin \omega t}$$

Comparison of like terms on both sides yields

$$\begin{array}{lcl} D \sin \phi = A & \text{Divide the two equations} & \\ D \cos \phi = B & \Rightarrow & \tan \phi = \frac{A}{B} \end{array}$$

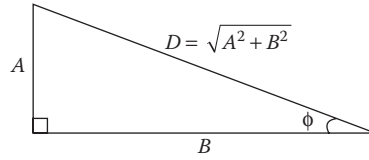


FIGURE 2.6
Phase shift.

Next, we construct a right triangle (Figure 2.6) in which angle ϕ satisfies $\tan \phi = A/B$:

$$\sin \phi = \frac{A}{D} = \frac{A}{\sqrt{A^2 + B^2}}, \quad \cos \phi = \frac{B}{D} = \frac{B}{\sqrt{A^2 + B^2}}$$

Since $D > 0$ always, depending on the signs of A and B , the proper quadrant and measure for ϕ can be readily determined. In summary,

$$A \cos \omega t + B \sin \omega t = D \sin(\omega t + \phi), \quad D = \sqrt{A^2 + B^2}, \quad \phi = \tan^{-1} \frac{A}{B} \quad (2.14)$$

Example 2.10: Amplitude, Phase Shift

Express $\sin t - 2 \cos t$ as $D \sin(\omega t + \phi)$.

Solution

Comparing with the general form $A \cos \omega t + B \sin \omega t$, we have $A = -2$, $B = 1$, and the frequency $\omega = 1$ rad/s. We also note that $D = \sqrt{5} > 0$. Since $\sin \phi = A/D < 0$ and $\cos \phi = B/D > 0$, we conclude that ϕ is in the fourth quadrant and $\phi = \tan^{-1}(-2) = -1.1071$ rad. In summary,

$$\sin t - 2 \cos t = \sqrt{5} \sin(t - 1.1071)$$

PROBLEM SET 2.2

In Problems 1 through 4, solve the linear, first-order IVP.

1. $\dot{x} + \frac{1}{3}x = \cos 2t$, $x(0) = 1$
2. $\frac{1}{2}\dot{x} + tx = \frac{1}{2}t$, $x(0) = \frac{1}{3}$
3. $(t-1)\dot{u} + tu = 2t$, $u(0) = 1$
4. $\dot{u} = (1-u)\cos t$, $u(0) = \frac{1}{2}$

In Problems 5 through 10, solve the linear, second-order IVP.

5. $\ddot{x} + 4\dot{x} + 4x = e^{-t}$, $x(0) = 1$, $\dot{x}(0) = 1$
6. $\ddot{x} + 4x = t$, $x(0) = 0$, $\dot{x}(0) = 1$
7. $\ddot{x} + x = \sin t$, $x(0) = 1$, $\dot{x}(0) = 0$
8. $\ddot{x} + 4\dot{x} + 3x = 2e^{-3t}$, $x(0) = 0$, $\dot{x}(0) = -1$
9. $6\ddot{u} + 7\dot{u} + 2u = 65 \cos t$, $u(0) = 0$, $\dot{u}(0) = 5$
10. $4\ddot{u} + 4\dot{u} + 5u = 0$, $u(0) = 0$, $\dot{u}(0) = 1$

In Problems 11 through 14, write the expression in the form $D\sin(\omega t + \phi)$.

11. $\cos t + \frac{1}{3}\sin t$
12. $\frac{1}{2}\cos 3t - \sin 3t$
13. $\cos(2t + \frac{1}{3}\pi)$
14. $3\sin \omega t - \cos \omega t$

In Problems 15 and 16, write the expression in the form $D\cos(\omega t + \phi)$.

15. $\cos t - \frac{3}{4}\sin t$
16. $\sin t - \cos t$

2.3 Laplace Transformation

In [Section 2.2](#), we learned that ODEs with constant coefficients may be solved entirely in time domain by using the method of undetermined coefficients, which has limitations in its applications. In this section, we introduce an alternative and systematic approach to solve these equations, with the advantages that a wider range of forcing functions can be handled and the arbitrary constants in a general solution do not need to be obtained separately. The idea is to transform an IVP via Laplace transformation to the s -domain, in which the transformed problem is algebraic and thus much easier to solve. This algebraic problem is then properly treated, and the data are ultimately transformed back to time domain to find the solution of the original IVP. If a function $x(t)$ is defined for all $t \geq 0$, its Laplace transform is defined by

$$X(s) \stackrel{\text{Notation}}{=} \mathcal{L}\{x(t)\} \stackrel{\text{Definition}}{=} \int_0^{\infty} e^{-st} x(t) dt \quad (2.15)$$

provided that the integral exists. The complex variable s is the Laplace variable, and $\mathcal{L}$ is the Laplace transform operator. It is a common practice to denote a time-dependent function by a lowercase letter such as $x(t)$ and its Laplace transform by the same letter in capital, $X(s)$. Transforming the data back to the time domain is done through inverse Laplace transformation $\mathcal{L}^{-1}$ as ([Figure 2.7](#))

$$x(t) = \mathcal{L}^{-1}\{X(s)\}$$

Laplace transforms of several functions are listed in [Table 2.2](#).

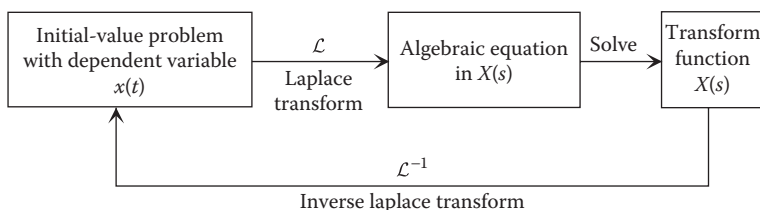


FIGURE 2.7

Solution process for initial-value problems, using Laplace transformation.

TABLE 2.2

Laplace Transform Pairs

No.	$f(t)$	$F(s)$
1	Unit impulse $\delta(t)$	1
2	1, unit step $u_s(t)$	$1/s$
3	t , unit ramp $u_r(t)$	$1/s^2$
4	$\delta(t-a)$	e^{-as}
5	$u(t-a)$	e^{-as}/s
6	t^{n-1} , $n = 1, 2, \dots$	$(n-1)!/s^n$
7	t^{a-1} , $a > 0$	$\Gamma(a)^*/s^a$
8	e^{-at}	$\frac{1}{s+a}$
9	te^{-at}	$\frac{1}{(s+a)^2}$
10	$t^n e^{-at}$, $n = 1, 2, \dots$	$\frac{n!}{(s+a)^{n+1}}$
11	$\frac{1}{b-a}(e^{-at} - e^{-bt})$, $a \neq b$	$\frac{1}{(s+a)(s+b)}$
12	$\frac{1}{a-b}(ae^{-at} - be^{-bt})$, $a \neq b$	$\frac{s}{(s+a)(s+b)}$
13	$\frac{1}{ab} \left[1 + \frac{1}{a-b}(be^{-at} - ae^{-bt}) \right]$	$\frac{1}{s(s+a)(s+b)}$
14	$\frac{1}{a^2}(-1 + at + e^{-at})$	$\frac{1}{s^2(s+a)}$
15	$\frac{1}{a^2}(1 - e^{-at} - ate^{-at})$	$\frac{1}{s(s+a)^2}$
16	$\sin \omega t$	$\frac{\omega}{s^2 + \omega^2}$
17	$\cos \omega t$	$\frac{s}{s^2 + \omega^2}$
18	$e^{-\sigma t} \sin \omega t$	$\frac{\omega}{(s+\sigma)^2 + \omega^2}$
19	$e^{-\sigma t} \cos \omega t$	$\frac{s+\sigma}{(s+\sigma)^2 + \omega^2}$
20	$1 - \cos \omega t$	$\frac{\omega^2}{s(s^2 + \omega^2)}$
21	$\omega t - \sin \omega t$	$\frac{\omega^3}{s^2(s^2 + \omega^2)}$
22	$t \cos \omega t$	$\frac{s^2 - \omega^2}{(s^2 + \omega^2)^2}$
23	$\frac{1}{2\omega} t \sin \omega t$	$\frac{s}{(s^2 + \omega^2)^2}$

(Continued)

TABLE 2.2 (Continued)

Laplace Transform Pairs

No.	$f(t)$	$F(s)$
24	$\frac{1}{2\omega^3}(\sin \omega t - \omega t \cos \omega t)$	$\frac{1}{(s^2 + \omega^2)^2}$
25	$\frac{1}{2\omega}(\sin \omega t + \omega t \cos \omega t)$	$\frac{s^2}{(s^2 + \omega^2)^2}$
26	$\frac{1}{\omega_2^2 - \omega_1^2} \left[\frac{1}{\omega_2} \sin \omega_2 t - \frac{1}{\omega_1} \sin \omega_1 t \right], \omega_1^2 \neq \omega_2^2$	$\frac{1}{(s^2 + \omega_1^2)(s^2 + \omega_2^2)}$
27	$\frac{1}{\omega_2^2 - \omega_1^2}(\cos \omega_1 t - \cos \omega_2 t), \omega_1^2 \neq \omega_2^2$	$\frac{s}{(s^2 + \omega_1^2)(s^2 + \omega_2^2)}$
28	$\sinh at$	$\frac{a}{s^2 - a^2}$
29	$\cosh at$	$\frac{s}{s^2 - a^2}$
30	$\frac{1}{a^2 - b^2} \left[\frac{1}{a} \sinh at - \frac{1}{b} \sinh bt \right], a \neq b$	$\frac{1}{(s^2 - a^2)(s^2 - b^2)}$
31	$\frac{1}{a^2 - b^2} [\cosh at - \cosh bt], a \neq b$	$\frac{s}{(s^2 - a^2)(s^2 - b^2)}$
32	$\frac{1}{3a^2} [e^{-at} + 2e^{\frac{1}{2}at} \sin(\frac{\sqrt{3}}{2}at - \frac{1}{6}\pi)]$	$\frac{1}{s^3 + a^3}$
33	$\frac{1}{3a} [-e^{-at} + 2e^{\frac{1}{2}at} \sin(\frac{\sqrt{3}}{2}at + \frac{1}{6}\pi)]$	$\frac{s}{s^3 + a^3}$
34	$\frac{1}{3a^2} [e^{at} - 2e^{-\frac{1}{2}at} \sin(\frac{\sqrt{3}}{2}at + \frac{1}{6}\pi)]$	$\frac{1}{s^3 - a^3}$
35	$\frac{1}{3a} [e^{-at} + 2e^{-\frac{1}{2}at} \sin(\frac{\sqrt{3}}{2}at - \frac{1}{6}\pi)]$	$\frac{s}{s^3 - a^3}$
36	$\frac{1}{4a^3} [\cosh at \sin at - \sinh at \cos at]$	$\frac{1}{s^4 + 4a^4}$
37	$\frac{1}{2a^2} \sinh at \sin at$	$\frac{s}{s^4 + 4a^4}$
38	$\frac{1}{2a^3} (\sinh at - \sin at)$	$\frac{1}{s^4 - a^4}$
39	$\frac{1}{2a^2} (\cosh at - \cos at)$	$\frac{s}{s^4 - a^4}$

*Gamma function $\Gamma(a) = \int_0^{\infty} t^{a-1} e^{-t} dt$.

Example 2.11: Laplace Transform

- Find $\mathcal{L}\{e^{-at}\}$, $a = \text{const.}$
-  Repeat (a) in MATLAB.

Solution

$$\text{a. } \mathcal{L}\{e^{-at}\} = \int_0^{\infty} e^{-st} e^{-at} dt = \int_0^{\infty} e^{-(s+a)t} dt = \frac{1}{s+a}$$

b. 

```
>> syms a t
>> F = laplace(exp(-a*t))
F =
1/(a + s)
```

2.3.1 Linearity of Laplace and Inverse Laplace Transforms

The Laplace transform operator $\mathcal{L}$ is linear; that is, if the Laplace transforms of functions $x_1(t)$ and $x_2(t)$ exist, and c_1 and c_2 are constant scalars, then

$$\begin{aligned}\mathcal{L}\{c_1x_1(t) + c_2x_2(t)\} &= \int_0^{\infty} e^{-st} [c_1x_1(t) + c_2x_2(t)] dt = c_1 \int_0^{\infty} e^{-st} x_1(t) dt + c_2 \int_0^{\infty} e^{-st} x_2(t) dt \\ &= c_1 \mathcal{L}\{x_1(t)\} + c_2 \mathcal{L}\{x_2(t)\} = c_1 X_1(s) + c_2 X_2(s)\end{aligned}$$

To establish the linearity of $\mathcal{L}^{-1}$, take the inverse Laplace transforms of the expressions on the far left and far right of the above-mentioned equation to obtain

$$c_1x_1(t) + c_2x_2(t) = \mathcal{L}^{-1}\{c_1X_1(s) + c_2X_2(s)\}$$

Noting that $x_1(t) = \mathcal{L}^{-1}\{X_1(s)\}$ and $x_2(t) = \mathcal{L}^{-1}\{X_2(s)\}$, the result follows.

2.3.2 Differentiation and Integration of Laplace Transforms

Let us consider two specific types of situations: $\mathcal{L}\{tg(t)\}$ and $\mathcal{L}\{g(t)/t\}$. In both cases, it is assumed that $G(s) = \mathcal{L}\{g(t)\}$ is either known directly from [Table 2.2](#) or can be determined by other means. Either way, once $G(s)$ is available, the two above-mentioned transforms will be obtained in terms of the derivative and the integral of $G(s)$, respectively. We first make the following definition. If a transform function is in the form $G(s) = N(s)/D(s)$, then any value of s for which $D(s) = 0$ is called a pole of $G(s)$. A pole with a *multiplicity* (number of occurrences) of one is known as a simple pole.

2.3.2.1 Differentiation of Laplace Transforms

If $G(s) = \mathcal{L}\{g(t)\}$ exists, then at any point except at the poles of $G(s)$, we have

$$\mathcal{L}\{tg(t)\} = -\frac{dG(s)}{ds} \quad (2.16)$$

In the general case,

$$\mathcal{L}\{t^n g(t)\} = (-1)^n \frac{d^n G(s)}{ds^n}$$

Example 2.12: Derivative of Laplace Transforms

- a. Find $\mathcal{L}\{te^{-2t/3}\}$.
 b.  Repeat (a) in MATLAB.

Solution

- a. Comparison with Equation 2.16 reveals $g(t) = e^{-2t/3}$, so that $G(s) = \frac{1}{s + \frac{2}{3}}$. Then,

$$\mathcal{L}\{te^{-2t/3}\} = -\frac{d}{ds}\left(\frac{1}{s + \frac{2}{3}}\right) = \left(\frac{1}{s + \frac{2}{3}}\right)^2$$

- b. 

```
>> syms t
>> laplace(t*exp(-2*t/3))
ans =
1/(s + 2/3)^2
```

2.3.2.2 Integration of Laplace Transforms

If $\mathcal{L}\{g(t)/t\}$ exists and the order of integration can be interchanged, then

$$\mathcal{L}\left\{\frac{g(t)}{t}\right\} = \int_s^\infty G(\sigma) d\sigma \quad (2.17)$$

Example 2.13: Integral of Laplace Transforms

- a. Find $\mathcal{L}\left\{\frac{\sin \omega t}{t}\right\}$.
 b.  Repeat (a) in MATLAB.

Solution

- a. Comparing with Equation 2.17, we find $g(t) = \sin \omega t$, so that $G(s) = \frac{\omega}{s^2 + \omega^2}$. Consequently,

$$\mathcal{L}\left\{\frac{\sin \omega t}{t}\right\} = \int_s^\infty \frac{\omega}{\sigma^2 + \omega^2} d\sigma = \int_s^\infty \frac{1}{1 + (\sigma/\omega)^2} \frac{d\sigma}{\omega} = \left[\tan^{-1} \frac{\sigma}{\omega} \right]_{\sigma=s}^\infty = \frac{\pi}{2} - \tan^{-1} \frac{s}{\omega} = \tan^{-1} \frac{\omega}{s}$$

- b. 

```
>> syms w t
>> laplace(sin(w*t)/t)
ans =
atan(w/s)
```

2.3.3 Special Functions

Characteristics of dynamic systems, such as stability and damping levels, can be largely determined by studying their responses to external excitations or disturbances, which are mathematically modeled as special functions. In this section, we introduce the step, ramp, pulse, and impulse functions, as well as their Laplace transforms.

2.3.3.1 Unit-Step Function

The unit-step function occurring at $t = 0$ is shown in [Figure 2.8a](#) and is defined as

$$u(t) = \begin{cases} 1 & \text{if } t > 0 \\ 0 & \text{if } t < 0 \\ \text{undefined (finite)} & \text{if } t = 0 \end{cases}$$

Physically, this may be realized as a constant signal (of magnitude 1) suddenly applied to the system at time $t = 0$. Using the definition of the Laplace transform, Equation 2.15, we find

$$\mathcal{L}\{u(t)\} \stackrel{\text{Notation}}{=} U(s) = \int_0^{\infty} e^{-st} u(t) dt = \int_0^{\infty} e^{-st} dt = \frac{1}{s}$$

When the magnitude is A , the signal is called a step function, denoted by $Au(t)$, and

$$\mathcal{L}\{Au(t)\} = \int_0^{\infty} e^{-st} A dt = \frac{A}{s}$$

If the unit-step occurs at time $a \neq 0$ ([Figure 2.8b](#)), it is described by

$$u(t-a) = \begin{cases} 1 & \text{if } t > a \\ 0 & \text{if } t < a \\ \text{undefined (finite)} & \text{if } t = a \end{cases}$$

The time-delayed step of magnitude A is described by $Au(t-a)$ and is shown in [Figure 2.8c](#). In order to find the Laplace transform of this signal, we need the following shifting property:

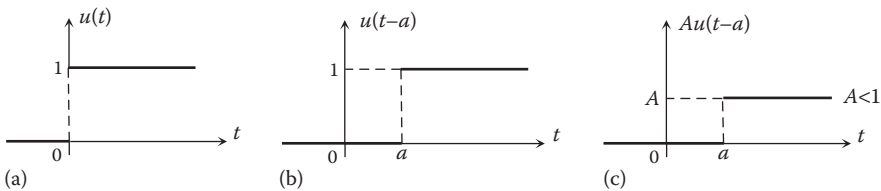


FIGURE 2.8

(a) Unit step at $t = 0$, (b) time-delayed unit step, and (c) time-delayed step.

Shift on t -axis: If $F(s) = \mathcal{L}\{f(t)\}$ exists and $a \geq 0$, then

$$\mathcal{L}\{f(t-a)u(t-a)\} = e^{-as}F(s) \quad (2.18)$$

Comparing $\mathcal{L}\{Au(t-a)\}$ to the left side of Equation 2.18, we find $f(t-a) = A$, which in turn implies $f(t) = A$ and hence $F(s) = A/s$. Therefore,

$$\mathcal{L}\{Au(t-a)\} = \frac{Ae^{-as}}{s}$$

Example 2.14: Shift on t -Axis

Consider the function $g(t)$ defined in Figure 2.9.

- Express $g(t)$ in terms of unit-step functions.
- Find $G(s)$ by using the shift on t -axis, Equation 2.18.
- Confirm the result of (b) in MATLAB.

Solution

- We construct the signal in Figure 2.9 by subtracting Figure 2.10b from 2.10a, that is,

$$g(t) = \frac{3}{2}tu(t) - \frac{3}{2}tu(t-1)$$

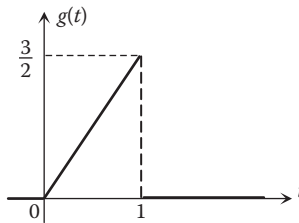


FIGURE 2.9
Example 2.14.

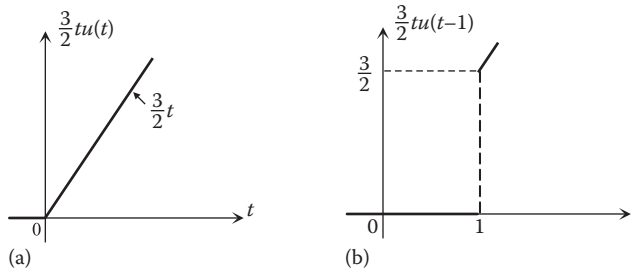


FIGURE 2.10
(a) Infinite ray and (b) tail end of the ray.

b. Applying Equation 2.18 to the two terms yields

$$\mathcal{L}\left\{\frac{3}{2}tu(t)\right\} = \frac{3}{s^2}, \quad \mathcal{L}\left\{\frac{3}{2}tu(t-1)\right\} = \frac{3}{2}e^{-s}\left(\frac{1}{s^2} + \frac{1}{s}\right)$$

Therefore,

$$G(s) = \frac{3}{s^2} - \frac{3}{2}e^{-s}\left(\frac{1}{s^2} + \frac{1}{s}\right)$$

c. 

```
>> syms t
% heaviside represents the unit-step
>> g = (3/2)*t*heaviside(t) - (3/2)*t*heaviside(t-1);
>> G = laplace(g)
G =
3/(2*s^2) - (3*exp(-s))/(2*s^2) - (3*exp(-s))/(2*s)
```

2.3.3.2 Unit-Ramp Function

The unit-ramp function (Figure 2.11) is defined as

$$u_r(t) = \begin{cases} t & \text{if } t \geq 0 \\ 0 & \text{if } t < 0 \end{cases}$$

Physically, this models a signal that changes linearly with a unit rate. By Equation 2.15,

$$\mathcal{L}\{u_r(t)\} = U_r(s) = \int_0^{\infty} te^{-st} dt = \frac{1}{s^2}$$

In general, when the rate is A , the signal is called a ramp function, denoted by $Au_r(t)$, and

$$\mathcal{L}\{Au_r(t)\} = \frac{A}{s^2}$$

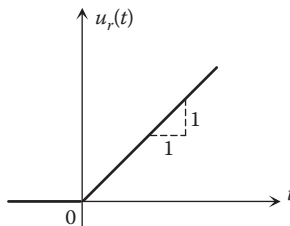


FIGURE 2.11
Unit ramp.

2.3.3.3 Unit-Pulse Function

The unit-pulse function (Figure 2.12a) is defined as

$$u_p(t) = \begin{cases} 1/t_1 & \text{if } 0 < t < t_1 \\ 0 & \text{if } t < 0 \text{ or } t > t_1 \end{cases}$$

The word “unit” signifies the unit area that the signal occupies. It is readily verified that

$$u_p(t) = \frac{1}{t_1} [u(t) - u(t - t_1)]$$

so that

$$\mathcal{L}\{u_p(t)\} = \frac{1}{t_1} \left[\frac{1}{s} - \frac{e^{-t_1 s}}{s} \right] = \frac{1 - e^{-t_1 s}}{t_1 s}$$

In general, if the signal occupies an area A , it is called a pulse function and is denoted by $Au_p(t)$.

2.3.3.4 Unit-Impulse (Dirac Delta) Function

In the unit-pulse function of Figure 2.12a, let $t_1 \rightarrow 0$ (Figure 2.12b). In the limit, the rectangular-shaped signal occupies a region with an infinitesimally small width and an infinitely large height (Figure 2.12c). However, the area remains unity throughout the process. This limiting signal is known as the unit-impulse (or Dirac delta) function, denoted by $\delta(t)$. In general, if the area is A , the signal is called an impulse, denoted by $A\delta(t)$. An impulse with zero duration and infinite magnitude is mathematical fabrication and does not occur physically. However, if an external disturbance (such as an applied force, voltage, or torque) is a pulse with a very large magnitude and is applied for a very short period of time, then it can be approximated as an impulse. Since $\delta(t)$ is the limit of $u_p(t)$ as $t_1 \rightarrow 0$, we have

$$\mathcal{L}\{\delta(t)\} \stackrel{\text{Notation}}{=} \Delta(s) = \lim_{t_1 \rightarrow 0} \left\{ \frac{1 - e^{-st_1}}{st_1} \right\} \stackrel{\text{L'Hôpital's rule}}{=} \lim_{t_1 \rightarrow 0} \left\{ \frac{se^{-st_1}}{s} \right\} = 1$$

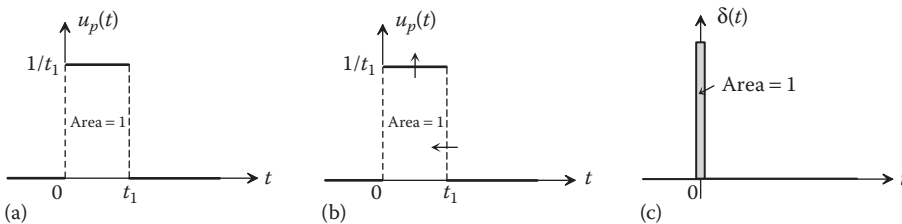


FIGURE 2.12

(a) Unit pulse, (b) unit pulse as $t_1 \rightarrow 0$, and (c) unit impulse.

If the unit-impulse occurs at $t = a$, it is represented by $\delta(t - a)$ and has the properties

$$\delta(t - a) = \begin{cases} 0 & \text{if } t \neq a \\ \infty & \text{if } t = a \end{cases}$$

and

$$\int_{-\infty}^{\infty} \delta(t - a) dt = 1$$

It can also be shown that

$$\mathcal{L}\{\delta(t - a)\} = e^{-as}$$

2.3.3.5 The Relation between Unit-Impulse and Unit-Step Functions

Noting that $t = a$ is a point of discontinuity of $u(t - a)$, the unit-impulse signal $\delta(t - a)$ can be regarded as the derivative of $u(t - a)$ at this point of discontinuity, that is,

$$\frac{d}{dt} u(t - a) = \delta(t - a)$$

Therefore, the idea of the impulse function allows for the differentiation of time-variant functions with discontinuities.

2.3.3.6 Periodic Functions

Dynamic systems are often subjected to excitations that exhibit repeated behavior over long periods of time. Such excitations are modeled as periodic functions. A function $f(t)$ is said to be periodic with period $P > 0$ if it is defined for all $t > 0$ and $f(t + P) = f(t)$ for all $t > 0$. We assume that $f(t)$ is also piecewise continuous, as basically all signals of physical interest are. In order to determine a system's response to a specific input, the Laplace transform of the input must be identified properly. The Laplace transform of a periodic function $f(t)$ of period P is found as follows. By Equation 2.15,

$$F(s) = \int_0^{\infty} e^{-s\tau} f(t) dt = \int_0^P e^{-s\tau} f(t) dt + \int_P^{2P} e^{-s\tau} f(t) dt + \dots = \sum_{k=0}^{\infty} \left\{ \int_{kP}^{(k+1)P} e^{-s\tau} f(t) dt \right\}$$

To make the lower and upper limits of the integral independent of P , we introduce the dummy variable $\tau = t - kP$. Then,

$$F(s) = \sum_{k=0}^{\infty} \left\{ \int_0^P e^{-s(\tau+kP)} f(\tau) d\tau \right\} = \sum_{k=0}^{\infty} e^{-skP} \left\{ \int_0^P e^{-s\tau} f(\tau) d\tau \right\}$$

Note that the summation only affects the exponential term because the integral term is independent of the index k . However, the exponential term is a geometric series, so that

$$\sum_{k=0}^{\infty} e^{-skP} = \sum_{k=0}^{\infty} (e^{-sP})^k \stackrel{\text{Geometric series}}{=} \frac{1}{1 - e^{-sP}}$$

Using this information in the expression of $F(s)$, we find

$$F(s) = \left\{ \int_0^P e^{-s\tau} f(\tau) d\tau \right\} \sum_{k=0}^{\infty} e^{-skP} = \left\{ \int_0^P e^{-s\tau} f(\tau) d\tau \right\} \frac{1}{1 - e^{-sP}}$$

Therefore, the Laplace transform of the periodic function is determined as

$$F(s) = \frac{1}{1 - e^{-Ps}} \int_0^P e^{-st} f(t) dt \quad (2.19)$$

Example 2.15: Periodic Signal

Find the Laplace transform of the periodic function $g(t)$, defined in [Figure 2.13](#).

Solution

It is clear that the period is $P=1$. Noting that $g(t)=t$ for $0 < t < 1$, the integral in Equation 2.19 is evaluated as

$$\int_0^1 e^{-st} g(t) dt = \int_0^1 e^{-st} t dt = \frac{1 - e^{-s}(s+1)}{s^2}$$

Then, by Equation 2.19,

$$G(s) = \frac{1 - e^{-s}(s+1)}{s^2(1 - e^{-s})}$$

2.3.4 Laplace Transforms of Derivatives and Integrals

Mathematical models of dynamic systems generally involve differential equations of various orders. In some occasions, the governing equation(s) may contain derivatives,

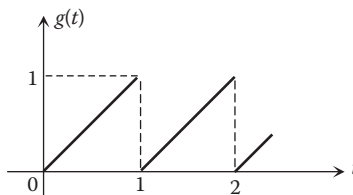


FIGURE 2.13
Periodic function in Example 2.15.

as well as integrals of a function. For example, an electrical circuit with a capacitor falls in this category of systems. Consequently, using the Laplace transform approach to solve such equations requires knowledge of the Laplace transform of derivatives and integrals.

2.3.4.1 Laplace Transforms of Derivatives

The Laplace transform of the n th-order derivative of $x(t)$ is given by

$$\mathcal{L}\{x^{(n)}(t)\} = s^n X(s) - s^{n-1}x(0) - s^{n-2}\dot{x}(0) - \dots - x^{(n-1)}(0)$$

In particular, for the first and second derivatives, the above-mentioned yields

$$\begin{aligned}\mathcal{L}\{\dot{x}(t)\} &= sX(s) - x(0) \\ \mathcal{L}\{\ddot{x}(t)\} &= s^2X(s) - sx(0) - \dot{x}(0)\end{aligned}\tag{2.20}$$

These relations are directly employed in solving IVPs.

2.3.4.2 Laplace Transforms of Integrals

The Laplace transform of the integral of a function $x(t)$ is given by

$$\mathcal{L}\left\{\int_0^t x(t)dt\right\} = \frac{1}{s}X(s)\tag{2.21}$$

Equivalently,

$$\mathcal{L}^{-1}\left\{\frac{1}{s}X(s)\right\} = \int_0^t x(t)dt\tag{2.22}$$

Example 2.16: Inverse Laplace Transform

- Find $\mathcal{L}^{-1}\left\{\frac{1}{s(s+\frac{1}{2})}\right\}$.
-  Confirm the result of (a) in MATLAB.

Solution

- Comparing with the left side of Equation 2.22, $X(s) = \frac{1}{s+\frac{1}{2}}$, so that $x(t) = e^{-t/2}$.
Then,

$$\mathcal{L}^{-1}\left\{\frac{1}{s(s+\frac{1}{2})}\right\} = \int_0^t e^{-t/2}dt = 2(1 - e^{-t/2})$$

b. 

```

>> syms s
>> ilaplace(1/s/(s+1/2))
ans =
2 - 2*exp(-t/2)
% Performing the integration in (a) yields the same result.
>> syms t
>> int(exp(-t/2), 0, t)
ans =
2 - 2*exp(-t/2) % Result confirmed

```

2.3.5 Inverse Laplace Transformation

As was shown in Figure 2.7, the final step in solving an IVP involves the inverse Laplace transform $\mathcal{L}^{-1}\{X(s)\}$. This can essentially be done using one of two techniques: (1) partial-fraction expansion method and (2) convolution method.

2.3.5.1 Partial-Fraction Expansion Method

In system dynamics, the transform function $X(s)$ is normally in the form

$$X(s) = \frac{N(s)}{D(s)} = \frac{\text{Polynomial of degree } m}{\text{Polynomial of degree } n}, \quad m < n \quad (2.23)$$

The idea behind the partial-fraction expansion method is to express $X(s)$ as a sum of suitable fractions and find the inverse Laplace transform of each fraction, and ultimately, via the linearity of $\mathcal{L}^{-1}$, the sum of the resulting time functions yields $x(t)$. How these partial fractions are formed depends on the nature of the poles of $X(s)$, that is, the roots of $D(s)$, which may be real or complex.

- All roots of $D(s)$ are real

Suppose $D(s) = s^3 + 3s^2 + 2s$, so that its roots are $s = 0, -1, -2$ (all real). In this case, we express $D(s)$ as a product of linear factors, as $D(s) = s(s+1)(s+2)$.

- $D(s)$ has real and complex roots

Suppose $D(s) = s^3 + 2s^2 + 2s$, so that its roots are $s = 0, -1 \pm j$. In this case, instead of writing $D(s) = s(s+1+j)(s+1-j)$, we write $D(s) = s(s^2 + 2s + 2)$, with the second-degree polynomial with complex roots remaining intact. Any second-degree polynomial with complex roots is called an irreducible polynomial.

Regardless of whether poles of $X(s)$ are real or complex, four possible cases could arise.

Case (1): *Linear factor* $s - p_i$

If p_i is a simple pole of $X(s)$, then $D(s)$ contains the factor $s - p_i$. This factor is associated with a fraction in the form of

$$\frac{A}{s - p_i}$$

where $A = \text{const}$ is called a residue and is to be determined appropriately.

Example 2.17: Linear Factors

- a. Find $\mathcal{L}^{-1}\{X(s)\}$, where $X(s) = \frac{s+3}{(s+1)(s+2)}$.
 b.  Verify the result of (a) in MATLAB.

Solution

- a. $D(s)$ contains two linear factors, since poles of $X(s)$ are real and distinct: $-1, -2$. Each linear term is associated with a simple fraction, as mentioned earlier. Thus,

$$X(s) = \frac{s+3}{(s+1)(s+2)} = \boxed{\frac{A}{s+1}} + \boxed{\frac{B}{s+2}} = \frac{A(s+2) + B(s+1)}{(s+1)(s+2)} = \frac{(A+B)s + 2A + B}{(s+1)(s+2)}$$

The denominators of the original and the final fractions are identical (by design), so we force their respective numerators to be identical, that is,

$$s+3 \equiv (A+B)s + 2A + B$$

However, this identity holds only if the coefficients of like powers of s on both sides are the same. So, we have

$$\begin{array}{llll} \text{Coefficient of } s : & 1 = A + B & \begin{array}{c} \text{Solve} \\ \Rightarrow \end{array} & A = 2 \\ \text{Constant term :} & 3 = 2A + B & & B = -1 \end{array}$$

Insert the two residues into the partial fractions, and perform term-by-term inverse Laplace transformation to obtain

$$X(s) = \frac{2}{s+1} - \frac{1}{s+2} \xrightarrow{\mathcal{L}^{-1}} x(t) = 2\mathcal{L}^{-1}\left\{\frac{1}{s+1}\right\} - \mathcal{L}^{-1}\left\{\frac{1}{s+2}\right\} = 2e^{-t} - e^{-2t}$$

- b. 

```
>> syms s
>> x = ilaplace((s+3)/(s+1)/(s+2))
x =
2*exp(-t) - exp(-2*t)
```

Case (2): *Repeated linear factor* $(s-p_i)^k$

If p_i is a pole of $X(s)$ with multiplicity k , then $D(s)$ contains the factor $(s-p_i)^k$. This factor is then associated with partial fractions

$$\frac{A_k}{(s-p_i)^k} + \frac{A_{k-1}}{(s-p_i)^{k-1}} + \dots + \frac{A_2}{(s-p_i)^2} + \frac{A_1}{s-p_i}$$

where the residues $A_k, \dots, A_1$ are determined, as in Case (1).

Case (3): *Irreducible polynomial* $s^2 + as + b$

Each irreducible polynomial is associated with a single fraction in the form of

$$\frac{Bs+C}{s^2+as+b}$$

with constants B and C to be determined. Before taking the inverse Laplace transform, we complete the square in the irreducible polynomial, that is, $s^2 + as + b = (s + \sigma)^2 + \omega^2$. For example, $s^2 + 2s + 2 = (s + 1)^2 + 1^2$. Then, at some point, we need to determine

$$\mathcal{L}^{-1} \left\{ \frac{Bs + C}{(s + \sigma)^2 + \omega^2} \right\}$$

The key is to split the fraction in terms of the two expressions

$$\frac{\omega}{(s + \sigma)^2 + \omega^2}, \quad \frac{s + \sigma}{(s + \sigma)^2 + \omega^2}$$

so that we can ultimately use the relations (Table 2.2)

$$\mathcal{L}^{-1} \left\{ \frac{\omega}{(s + \sigma)^2 + \omega^2} \right\} = e^{-\sigma t} \sin \omega t, \quad \mathcal{L}^{-1} \left\{ \frac{s + \sigma}{(s + \sigma)^2 + \omega^2} \right\} = e^{-\sigma t} \cos \omega t$$

Example 2.18: Real and Complex Poles

- Find $\mathcal{L}^{-1}\{X(s)\}$, where $X(s) = \frac{4}{(s+3)(s^2+2s+5)}$.
-  Verify the result of (a) in MATLAB.

Solution

- The term $s^2 + 2s + 5 = (s + 1)^2 + 2^2$ is irreducible, whereas $s + 3$ is a linear factor. Therefore,

$$\begin{aligned} X(s) = \frac{4}{(s+3)(s^2+2s+5)} &= \boxed{\frac{A}{s+3}} + \boxed{\frac{Bs+C}{s^2+2s+5}} = \frac{A(s^2+2s+5) + (Bs+C)(s+3)}{(s+3)(s^2+2s+5)} \\ &= \frac{(A+B)s^2 + (2A+3B+C)s + 5A+3C}{(s+3)(s^2+2s+5)} \end{aligned}$$

Proceeding as before,

$$\begin{aligned} A+B &= 0 & A &= \frac{1}{2} \\ 2A+3B+C &= 0 & \Rightarrow B &= -\frac{1}{2} \\ 5A+3C &= 4 & C &= \frac{1}{2} \end{aligned}$$

Substituting into the partial fractions, we arrive at

$$X(s) = \frac{1}{2} \left\{ \frac{1}{s+3} - \frac{s-1}{s^2+2s+5} \right\} = \frac{1}{2} \left\{ \frac{1}{s+3} - \frac{s+1}{(s+1)^2+2^2} + \frac{2}{(s+1)^2+2^2} \right\}$$

Term-by-term Laplace inversion yields $x(t) = \frac{1}{2}[e^{-3t} - e^{-t} \cos 2t + e^{-t} \sin 2t]$.

- 

```
>> syms s
>> x = ilaplace(4/(s+3)/(s^2 + 2*s+5))
x =
exp(-3*t)/2 - (exp(-t)*(cos(2*t) - sin(2*t)))/2
```

Case (4): *Repeated irreducible polynomial* $(s^2 + as + b)^k$

Each factor $(s^2 + as + b)^k$ in $D(s)$ is associated with partial fractions

$$\frac{B_k s + C_k}{(s^2 + as + b)^k} + \dots + \frac{B_2 s + C_2}{(s^2 + as + b)^2} + \frac{B_1 s + C_1}{s^2 + as + b}$$

2.3.5.2 Performing Partial-Fraction Expansion in MATLAB

The command `ilaplace` returns the inverse Laplace transform but not the actual partial fractions. For that purpose, the `residue` command is used instead. This command is concerned with the partial-fraction expansion of the ratio of two polynomials in the form $B(s)/A(s)$, where, unlike the restriction cited in Equation 2.23, the degree of $B(s)$ can be higher than that of $A(s)$. When $\deg B(s) \geq \deg A(s)$, polynomial division results in

$$\frac{B(s)}{A(s)} = K(s) + \frac{R(1)}{s - P(1)} + \frac{R(2)}{s - P(2)} + \dots + \frac{R(n)}{s - P(n)} \quad (2.24)$$

where $K(s)$ is called the direct term and $R(1), \dots, R(n)$ are the residues. Note that the assumption here is that there are no multiple poles. For further information, type `help residue` at the MATLAB command prompt. The structure $[R, P, K] = \text{residue}(B, A)$ indicates that there are two input arguments (B, A) and three outputs $[R, P, K]$. Here, B and A are vectors of the coefficients of the numerator and denominator, respectively. The residues are returned in the column vector R , the pole locations in column vector P , and the direct terms in the row vector K .

Example 2.19: Revisiting Example 2.17

Find the partial-fraction expansion for $X(s) = \frac{s+3}{(s+1)(s+2)}$.

Solution

```
>> A = [1 3 2];      % denominator s^2+3*s+2
>> B = [1 3];        % numerator s+3
>> [R,P,K] = residue(B,A)
R =
    -1
     2          % residues are -1 and 2
P =
    -2
    -1          % poles (corresponding to residues) are -2 and -1
K =
    []          % empty set because deg B(s) < deg A(s)
```

Following the standard form in Equation 2.24, we find

$$X(s) = \frac{-1}{s - (-2)} + \frac{2}{s - (-1)} = \frac{-1}{s+2} + \frac{2}{s+1}$$

which completely agrees with the result of Example 2.17.

Example 2.20: Revisiting Example 2.18

Find the partial-fraction expansion for $X(s) = \frac{4}{(s+3)(s^2+2s+5)}$.

Solution

```
>> B = 4;
>> A = conv([1 3], [1 2 5])    % "conv" gives the product
A =
     1     5    11    15
>> [R,P,K] = residue(B,A)
R =
    0.5000 + 0.0000i
   -0.2500 - 0.2500i
   -0.2500 + 0.2500i

P =
   -3.0000 + 0.0000i
   -1.0000 + 2.0000i
   -1.0000 - 2.0000i

K =
     []
```

The first element in R is the residue associated with the first element (pole) in P. Therefore, the residue for the pole at -3 is 0.5 . The other two pairs of poles/residues are identified similarly.

$$X(s) = \frac{4}{(s+3)(s^2+2s+5)} = \frac{\frac{1}{2}}{s-(-3)} + \frac{-\frac{1}{4}-\frac{1}{4}j}{s-(-1+2j)} + \frac{-\frac{1}{4}+\frac{1}{4}j}{s-(-1-2j)}$$

Inspection confirms that the above is equivalent to the expansion obtained in Example 2.18.

2.3.5.3 Convolution Method

In systems analysis, the problem of determining the time history of a function often boils down to finding $\mathcal{L}^{-1}\{F(s)\}$, where $F(s) = G(s)H(s)$ and $g(t)$ and $h(t)$ are available. Then,

$$\mathcal{L}^{-1}\{F(s)\} = f(t) = \underbrace{(g * h)(t)}_{\text{Convolution of } g \text{ and } h} = \int_0^t g(\tau)h(t-\tau)d\tau \stackrel{\text{Symmetry}}{=} \int_0^t h(\tau)g(t-\tau)d\tau = (h * g)(t)$$

To prove this, it suffices to show

$$\mathcal{L}\left\{\int_0^t g(\tau)h(t-\tau)d\tau\right\} = G(s)H(s)$$

First, using the fact that $u(t-\tau) = 0$ for all $\tau > t$, the left side is rewritten as

$$\mathcal{L}\left\{\int_0^t g(\tau)h(t-\tau)d\tau\right\} = \mathcal{L}\left\{\int_0^\infty g(\tau)h(t-\tau)u(t-\tau)d\tau\right\}$$

Next, using Equation 2.15 and switching the order of integration, the above reduces to

$$\begin{aligned}\mathcal{L}\left\{\int_0^\infty g(\tau)h(t-\tau)u(t-\tau)d\tau\right\} &= \int_0^\infty e^{-st}\left[\int_0^\infty g(\tau)h(t-\tau)u(t-\tau)d\tau\right]dt \\ &= \int_0^\infty e^{-st}h(t-\tau)u(t-\tau)dt \int_0^\infty g(\tau)d\tau\end{aligned}$$

This is allowed because the Laplace transforms of g and h exist. Once again, since $u(t-\tau) = 0$ for all $\tau > t$, the first integral can be rewritten to give

$$\mathcal{L}\left\{\int_0^\infty g(\tau)h(t-\tau)u(t-\tau)d\tau\right\} = \int_\tau^\infty e^{-st}h(t-\tau)dt \int_0^\infty g(\tau)d\tau$$

Introducing the change of variables $\xi = t - \tau$, the above becomes

$$\begin{aligned}\mathcal{L}\left\{\int_0^\infty g(\tau)h(t-\tau)u(t-\tau)d\tau\right\} &= \int_0^\infty e^{-s(\xi+\tau)}h(\xi)d\xi \int_0^\infty g(\tau)d\tau \\ &= \underbrace{\int_0^\infty e^{-s\xi}h(\xi)d\xi}_{H(s)} \underbrace{\int_0^\infty e^{-s\tau}g(\tau)d\tau}_{G(s)} = H(s)G(s)\end{aligned}$$

Example 2.21: Convolution

Find $\mathcal{L}^{-1}\left\{\frac{1}{s^2(2s+1)}\right\}$ by means of convolution.

Solution

Let $F(s) = \frac{1}{s^2(2s+1)} = G(s)H(s)$, where $G(s) = \frac{1}{s^2}$ and $H(s) = \frac{1}{2s+1}$. Noting $g(t) = t$ and $h(t) = \frac{1}{2}e^{-t/2}$, we find

$$f(t) = (g * h)(t) = \int_0^t \tau \cdot \frac{1}{2} e^{-(t-\tau)/2} d\tau \stackrel{\text{Integration by parts}}{=} \left[\tau e^{-(t-\tau)/2} \right]_{\tau=0}^t - \int_0^t e^{-(t-\tau)/2} d\tau = t - 2 + 2e^{-t/2}$$

2.3.5.3.1 Solving Initial-Value Problems

The foregoing analysis has provided us with the necessary tools to complete the procedure of solving IVPs, as described in [Figure 2.7](#). This will be illustrated by the following examples.

Example 2.22: Second-Order Initial-Value Problem

Solve $\ddot{x} + 4\dot{x} + 3x = 0$, $x(0) = 0$, $\dot{x}(0) = \frac{1}{3}$.

Solution

Laplace transform of the ODE yields

$$s^2 X(s) - sx(0) - \dot{x}(0) + 4[sX(s) - x(0)] + 3X(s) = 0$$

Using the initial conditions and solving for $X(s)$, we find

$$X(s) = \frac{\frac{1}{3}}{(s+1)(s+3)} \xrightarrow[\text{or convolution}]{\substack{\mathcal{L}^{-1} \text{ using partial-fraction expansion} \\ \Rightarrow}} x(t) = \frac{1}{6}(e^{-t} - e^{-3t})$$

Example 2.23: Second-Order Initial-Value Problem

- Solve $\ddot{x} + 4x = 4u(t-1)$, $x(0) = 0$, $\dot{x}(0) = 0$, where $u(t)$ is the unit-step function.
-  Repeat (a) in MATLAB.

Solution

- Laplace transformation of the ODE, using Equation 2.20, taking into account the zero initial conditions, we arrive at

$$(s^2 + 4)X(s) = \frac{4}{s}e^{-s} \Rightarrow X(s) = \frac{4}{s(s^2 + 4)}e^{-s}$$

Let $H(s) = \frac{4}{s(s^2 + 4)}$, so that $X(s) = H(s)e^{-s}$. Using the shift on t -axis, Equation 2.18,

$$x(t) = \mathcal{L}^{-1}\{H(s)e^{-s}\} = h(t-1)u(t-1)$$

To find $h(t) = \mathcal{L}^{-1}\{H(s)\}$, we proceed as follows:

$$H(s) = \frac{4}{s(s^2 + 4)} \stackrel{\text{Partial fractions}}{=} \frac{A}{s} + \frac{Bs + C}{s^2 + 4} = \frac{(A+B)s^2 + Cs + 4A}{s(s^2 + 4)}$$

Equating the coefficients of like powers of s on both sides yields $A = 1$, $B = -1$, and $C = 0$. Then,

$$H(s) = \frac{1}{s} + \frac{-s}{s^2 + 4} \Rightarrow h(t) = 1 - \cos 2t$$

Finally, $x(t) = h(t-1)u(t-1) = [1 - \cos 2(t-1)]u(t-1)$. Convolution method would have led to the same result.

- 

```
>> x = simple(dsolve('D2x + 4*x = 4*heaviside(t-1)', 'x(0) = 0,
Dx(0) = 0'))
x =
-heaviside(t - 1)*(cos(2*t - 2) - 1)
```

This completely agrees with the result in (a).

2.3.6 Final-Value Theorem and Initial-Value Theorem

If a function $x(t)$ attains a finite limit as $t \rightarrow \infty$, that is, if it settles down after a sufficiently long time, then the limit is the steady-state value (or final value) of $x(t)$, denoted by x_{ss} . In other words, $x_{ss} = \lim_{t \rightarrow \infty} x(t)$. In many situations, the time history $x(t)$ is not available, but instead, the Laplace transform $X(s)$ is available. The final-value theorem (FVT) allows us to find x_{ss} , when it exists, by directly using $X(s)$ without knowledge of $x(t)$.

2.3.6.1 Final-Value Theorem

Suppose $X(s)$ has no poles in the right half plane (RHP) or on the imaginary axis, except possibly a simple pole (multiplicity of 1) at the origin. Then, $x(t)$ has a definite steady-state value, and it is given by

$$x_{ss} = \lim_{s \rightarrow 0} \{sX(s)\} \quad (2.25)$$

The FVT must be used only when it is applicable. Using the FVT when the required conditions do not hold may yield incorrect results.

Example 2.24: Final-Value Theorem

- Find x_{ss} if $X(s) = \frac{2s+1}{s(s^2+4s+5)}$.
-  Confirm the result of (a) in MATLAB.

Solution

- The poles of $X(s)$ are at 0 and $-2 \pm j$. The complex conjugate pair lies in the left half plane, and 0 is a simple pole (at the origin), all allowed by the FVT. Therefore,

$$x_{ss} = \lim_{s \rightarrow 0} \{sX(s)\} = \lim_{s \rightarrow 0} \left\{ \frac{2s+1}{s^2+4s+5} \right\} = \frac{1}{5}$$

- 

```
>> syms s
>> X = (2*s+1)/s/(s^2+4*s+5);
>> xss = limit(s*X,s,0)
xss =
1/5
% Re-confirm the result by inspecting the limit of x(t) as t goes
to infinity
>> xt = ilaplace(X)
xt =
1/5 - (exp(-2*t))*(cos(t) - 8*sin(t))/5 % Time history x(t)
>> syms t
>> xss = limit(xt,t,inf)
xss =
1/5
```

Example 2.25: Final-Value Theorem Does Not Apply

Let $X(s) = \frac{2s}{s^2 + \frac{1}{4}}$, so that its poles are at $\pm \frac{1}{2}j$, on the imaginary axis, not permitted by the FVT. Therefore, FVT is not applicable and should not be applied. If it were to be applied, it would yield

$$x_{ss} = \lim_{s \rightarrow 0} \{sX(s)\} = \lim_{s \rightarrow 0} \left\{ \frac{2s^2}{s^2 + \frac{1}{4}} \right\} = 0$$

which is obviously false. To explain this, we first find $x(t) = \mathcal{L}^{-1}\{X(s)\} = 2\cos(\frac{1}{2}t)$. Then, it is clear that $\lim_{t \rightarrow \infty} x(t)$ does not exist, since $x(t)$ is oscillatory and there is no steady-state value.

2.3.6.2 Initial-Value Theorem

If $\lim_{s \rightarrow \infty} \{sX(s)\}$ exists, then the initial value of $x(t)$ is given by

$$x(0^+) = \lim_{s \rightarrow \infty} \{sX(s)\} \quad (2.26)$$

Note that in the case of the initial-value theorem (IVT), the poles of $X(s)$ are not limited to specific regions in the complex plane, as they were with the FVT.

Example 2.26: Initial-Value Theorem

If $X(s) = \frac{\frac{1}{3}s^2 + 1}{s^2(4s + 3)}$, find $x(0^+)$.

Solution

By Equation 2.26,

$$x(0^+) = \lim_{s \rightarrow \infty} \{sX(s)\} = \lim_{s \rightarrow \infty} \left\{ \frac{\frac{1}{3}s^2 + 1}{s(4s + 3)} \right\} = \frac{1}{12}$$

Of course, this may be readily verified in time domain as follows. First, the inverse Laplace transformation yields

$$x(t) = \mathcal{L}^{-1} \left\{ \frac{\frac{1}{3}s^2 + 1}{s^2(4s + 3)} \right\} = \frac{1}{3}t + \frac{19}{36}e^{-3t/4} - \frac{4}{9}$$

Then,

$$x(0^+) = \lim_{t \rightarrow 0} \left\{ \frac{1}{3}t + \frac{19}{36}e^{-3t/4} - \frac{4}{9} \right\} = \frac{1}{12}$$

In [Section 2.2](#), we mentioned that although the initial condition and the initial value of a quantity are generally the same, there are cases in systems analysis where the two values differ. These situations normally arise when there is an impulsive forcing function involved, as illustrated in the following example.

Example 2.27: Initial Condition $\neq$ Initial Value

Suppose the governing equation for a system is described by the IVP

$$2\ddot{x} + \dot{x} + 3x = \delta(t), \quad x(0^-) = 0, \quad \dot{x}(0^-) = 0$$

where $\delta(t)$ denotes the unit impulse. Recall that 0^- refers to the time immediately prior to $t = 0$, and $x(0^-)$ is the initial condition of x . Determine the initial values of x and $\dot{x}$, that is, $x(0^+)$ and $\dot{x}(0^+)$.

Solution

Taking the Laplace transform of the ODE, and using the zero initial conditions, we find

$$X(s) = \frac{1}{2s^2 + s + 3}$$

By the IVT,

$$x(0^+) = \lim_{s \rightarrow \infty} \{sX(s)\} = \lim_{s \rightarrow \infty} \left\{ \frac{s}{2s^2 + s + 3} \right\} = 0$$

Thus, $x(0^+) = x(0^-)$. To evaluate $\dot{x}(0^+)$, we apply the IVT while x is replaced with $\dot{x}$. In the meantime, note that $\mathcal{L}\{\dot{x}\} = sX(s) - x(0^-) = sX(s)$. Then,

$$\dot{x}(0^+) = \lim_{s \rightarrow \infty} \{s\mathcal{L}\{\dot{x}\}\} = \lim_{s \rightarrow \infty} \{s[sX(s)]\} = \lim_{s \rightarrow \infty} \left\{ \frac{s^2}{2s^2 + s + 3} \right\} = \frac{1}{2}$$

Therefore, $\dot{x}(0^+) \neq \dot{x}(0^-)$. The findings here are best understood when they are viewed from a physical standpoint. The governing differential equation in this example can realistically describe the motion of a mechanical system, where x and $\dot{x}$ denote the displacement and velocity, respectively, and the applied force is impulsive. The initial displacement clearly remains unchanged before and after the exertion of the impulsive force. However, since the velocity is the rate of change of displacement, it experiences a sudden change immediately after the force is applied. Figure 2.14 depicts the details related to the current example.

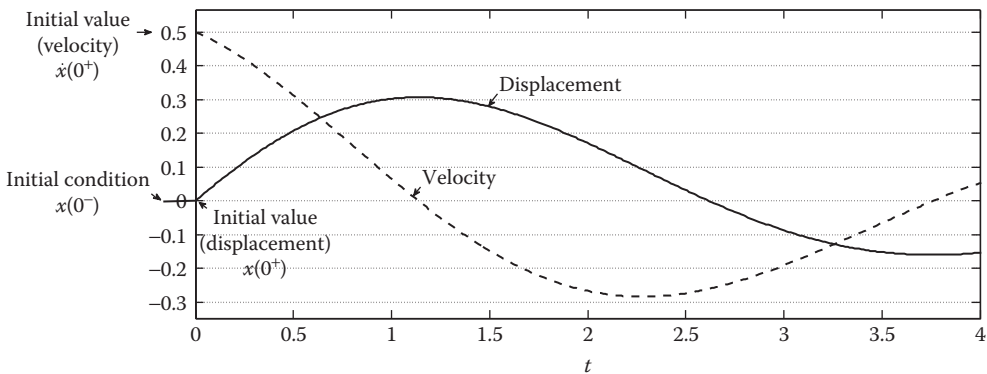


FIGURE 2.14
Relations between initial conditions and initial values.

PROBLEM SET 2.3

In Problems 1 through 8,

a. Find the Laplace transform of the given function. Use [Table 2.2](#) when applicable.

b.  Confirm the result of (a) in MATLAB.

1. e^{at+b} , $a, b = \text{const}$
2. $\cos(\omega t + \phi)$, $\omega, \phi = \text{const}$
3. $\sin(\omega t - \phi)$, $\omega, \phi = \text{const}$
4. $t^3 - \frac{1}{2}$
5. $\sin^2 t$
6. $t \cos \omega t$
7. $t \cosh t$
8. $t^2 \sin(\frac{1}{2}t)$

In Problems 9 through 12,

a. Express the signal in terms of unit-step functions.

b. Find the Laplace transform of the expression in (a) by using the shift on t -axis.

9. $g(t)$ in [Figure 2.15](#)

10. $g(t)$ in [Figure 2.16](#)

$$11. \quad g(t) = \begin{cases} 0 & \text{if } t < 0 \\ 1-t & \text{if } 0 < t < 1 \\ 0 & \text{if } t > 1 \end{cases}$$

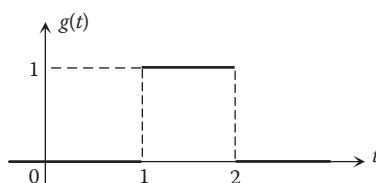


FIGURE 2.15

Signal in Problem 9.

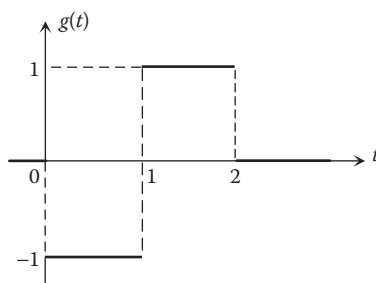


FIGURE 2.16

Signal in Problem 10

$$12. \quad g(t) = \begin{cases} 0 & \text{if } t < 0 \\ t & \text{if } 0 < t < 1 \\ 1 & \text{if } t > 1 \end{cases}$$

In Problems 13 through 16, find the Laplace transform of each periodic function whose definition in one period is given.

$$13. \quad h(t) = \begin{cases} 1 & \text{if } 0 < t < \frac{1}{2} \\ -1 & \text{if } \frac{1}{2} < t < 1 \end{cases}$$

$$14. \quad h(t) = \begin{cases} t & \text{if } 0 < t < 1 \\ 1-t & \text{if } 1 < t < 2 \end{cases}$$

$$15. \quad h(t) = \begin{cases} 1 & \text{if } 0 < t < 1 \\ 2-t & \text{if } 1 < t < 2 \end{cases}$$

$$16. \quad h(t) = 2-t, \quad 0 < t < 2$$

In Problems 17 and 18, find the Laplace transform of the given periodic function.

17. $f(t)$ in [Figure 2.17](#)

18. $f(t)$ in [Figure 2.18](#)

In Problems 19 and 20,

a. Find the partial-fraction expansion.

b.  Verify (a) by using the `residue` command in MATLAB.

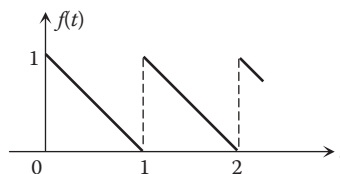


FIGURE 2.17

Periodic function in Problem 17.

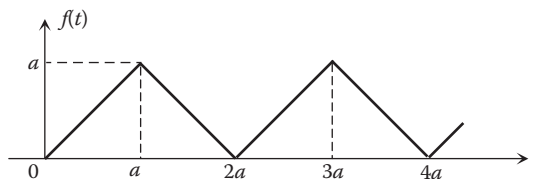


FIGURE 2.18

Periodic function in Problem 18.

$$19. \frac{s}{(s+1)(s^2+2s+2)}$$

$$20. \frac{9s+5}{s^2(s^2+4s+5)}$$

In Problems 21 through 26, find the inverse Laplace transform by using the partial-fraction expansion method.

$$21. \frac{s+1}{s^2(s+3)}$$

$$22. \frac{2s+3}{(s^2+4)(s+1)}$$

$$23. \frac{4s+5}{s^2(s^2+4s+5)}$$

$$24. \frac{s+10}{s(s^2+2s+5)}$$

$$25. \frac{5s+8}{s(s+2)^2}$$

$$26. \frac{2s^2+3s-4}{(s+3)(s^2+2s+2)}$$

In Problems 27 through 32,

a. Solve the IVP.

b.  Confirm the result of (a) in MATLAB.

$$27. \frac{2}{3}\dot{x} + 2x = u(t) - u(t-1), \quad x(0) = 0$$

$$28. \ddot{x} + 2\dot{x} + x = u(t-1) - u(t-2), \quad x(0) = 0, \quad \dot{x}(0) = 0$$

$$29. \ddot{x} + \frac{1}{4}\dot{x} = e^{-2t}, \quad x(0) = 0, \quad \dot{x}(0) = \frac{1}{2}$$

$$30. \ddot{x} + \dot{x} + \frac{1}{2}x = \delta(t-1), \quad x(0) = 0, \quad \dot{x}(0) = \frac{1}{2}$$

$$31. \ddot{x} + 3\dot{x} + 2x = \sin\left(\frac{2}{3}t\right), \quad x(0) = 0, \quad \dot{x}(0) = 3$$

$$32. \ddot{x} + 2\dot{x} + \frac{3}{4}x = e^{-t/2}, \quad x(0) = 0, \quad \dot{x}(0) = 0$$

In Problems 33 through 37, decide whether the FVT is applicable, and if so, find x_{ss} .

$$33. X(s) = \frac{s}{(4s+1)(s^2+3)}$$

$$34. X(s) = \frac{2s + \frac{1}{3}}{(s + \frac{1}{3})(s^2 + s + 1)}$$

$$35. X(s) = \frac{s+2}{(s+1)^3(s+3)}$$

$$36. X(s) = \frac{s+1}{s^2(s+3)(s+2)}$$

$$37. X(s) = \frac{4s^2+1}{s(4s^2+4s+2)}$$

In Problems 38 through 41, evaluate $x(0^+)$ by using the IVT.

$$38. \quad X(s) = \frac{3s^2 + 1}{(s+1)(\frac{2}{3}s+2)^2}$$

$$39. \quad X(s) = \frac{s^2 + 1}{(\frac{1}{2}s+1)(9s^2+6s+2)}$$

$$40. \quad X(s) = \frac{s(s+4)^2}{(s+1)(s+2)(s+3)^2}$$

$$41. \quad X(s) = \frac{\frac{1}{4}s^4 + 1}{s^3(2s^2+9)}$$

2.4 Summary

The rectangular (Cartesian) form of a complex number is $z = x + jy$, where x and y are the real and imaginary parts of z , respectively. The magnitude of z is $|z| = \sqrt{x^2 + y^2}$. The distance between two complex numbers z_1 and z_2 is $|z_1 - z_2|$. The complex conjugate of z , denoted by $\bar{z}$, is defined as $\bar{z} = x - jy$, and $z\bar{z} = x^2 + y^2 = |z|^2$. The polar form of z is $z = re^{j\theta}$, where $r = |z|$, and θ is measured from the positive real axis and regarded positive in the counterclockwise direction. Given a complex number $z = re^{j\theta} \neq 0$, and a positive integer n , the n th root of z is multi-valued and defined by

$$\sqrt[n]{z} = \sqrt[n]{r} \left(\cos \frac{\theta + 2k\pi}{n} + j \sin \frac{\theta + 2k\pi}{n} \right), \quad k = 0, 1, \dots, n-1$$

An n th-order ODE

$$a_n x^{(n)} + a_{n-1} x^{(n-1)} + \dots + a_1 \dot{x} + a_0 x = F(t)$$

is linear if coefficients $a_0, a_1, \dots, a_n$ are either constants or functions of the independent variable t . If $F(t) \equiv 0$, the ODE is homogeneous. Otherwise, it is nonhomogeneous. A linear, first-order ODE

$$\dot{x} + g(t)x = f(t)$$

has a general solution

$$x(t) = e^{-h} \left[\int e^h f(t) dt + c \right], \quad h = \int g(t) dt, \quad c = \text{const}$$

If a function $x(t)$ is defined for all $t \geq 0$, its Laplace transform is defined by

$$X(s) \stackrel{\text{Notation}}{=} \mathcal{L}\{x(t)\} \stackrel{\text{Definition}}{=} \int_0^{\infty} e^{-st} x(t) dt$$

provided that the integral exists. If $F(s) = \mathcal{L}\{f(t)\}$ exists and $a \geq 0$, then

$$\mathcal{L}\{f(t-a)u(t-a)\} = e^{-as}F(s)$$

The Laplace transform of a periodic function $f(t)$ with period P is given by

$$F(s) = \frac{1}{1-e^{-Ps}} \int_0^P e^{-st} f(t) dt$$

The Laplace transforms of the first and second derivatives of a function $x(t)$ are defined by

$$\mathcal{L}\{\dot{x}(t)\} = sX(s) - x(0)$$

$$\mathcal{L}\{\ddot{x}(t)\} = s^2X(s) - sx(0) - \dot{x}(0)$$

The Laplace transform of the integral of a function $x(t)$ is

$$\mathcal{L}\left\{\int_0^t x(t) dt\right\} = \frac{1}{s} X(s)$$

Inverse Laplace transformation can be performed by either partial-fraction expansion or the convolution method, which states

$$\mathcal{L}^{-1}\{G(s)H(s)\} = \underset{\text{Convolution of } g \text{ and } h}{(g * h)(t)} = \int_0^t g(\tau)h(t-\tau)d\tau \stackrel{\text{Symmetry}}{=} \int_0^t h(\tau)g(t-\tau)d\tau = (h * g)(t)$$

Final-value theorem: If $X(s)$ has no poles in the RHP or on the imaginary axis, except possibly a simple pole at the origin, then $x(t)$ has a definite steady-state value, and it is given by

$$x_{ss} = \lim_{s \rightarrow 0} \{sX(s)\}$$

Initial-value theorem: If $\lim_{s \rightarrow \infty} \{sX(s)\}$ exists, then the initial value of x is

$$x(0^+) = \lim_{s \rightarrow \infty} \{sX(s)\}$$

Review Problems

In Problems 1 through 4, perform the operations and express the result in rectangular form.

1. $e^{(-1+2j)/(1+j)}$
2. $e^{(0.15-0.85j)^8}$

$$3. \frac{(\frac{2}{3} + \frac{1}{2}j)^4}{(1 + \frac{3}{4}j)(-1 + \frac{1}{3}j)}$$

$$4. \frac{(1 - \frac{2}{3}j)^4}{(\frac{2}{3} + j)^3}$$

$$5. \text{Find all possible values of } (3 + \frac{1}{3}j)^{4/3}.$$

6. Find all possible roots of the following equation:

$$(2 - j)z^4 + 1 - 2j = 0$$

In Problems 7 through 10, solve the IVP. Do not use Laplace transformation.

$$7. \dot{y} - y \tan t = t + 1, \quad y(0) = 1$$

$$8. \dot{y} = (1 - y) \sin t, \quad y(\frac{1}{2}\pi) = 2$$

$$9. \ddot{y} - y = t^2 - \cos t, \quad y(0) = \frac{1}{2}, \quad \dot{y}(0) = 0$$

$$10. \ddot{y} + 4y = e^t \sin t, \quad y(0) = 0, \quad \dot{y}(0) = \frac{1}{10}$$

11. Express the signal in [Figure 2.19](#) in terms of unit-step functions and subsequently find its Laplace transform by using the shift on t -axis.

12. Find the Laplace transform of the periodic function in [Figure 2.20](#).

13. Find the response $x(t)$ of a system governed by

$$3\ddot{x} + 7\dot{x} + 2x = tu(t-1), \quad x(0) = 0, \quad \dot{x}(0) = 0$$

14. Evaluate the convolution $t * u(t-a)$.

15. Evaluate the convolution $(g * h)(t)$, where $g(t)$ is shown in [Figure 2.21](#) and $h(t) = \sin t$.

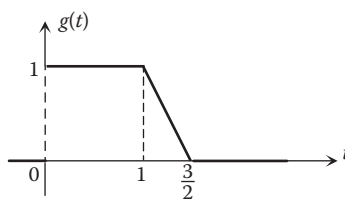


FIGURE 2.19
Problem 11.

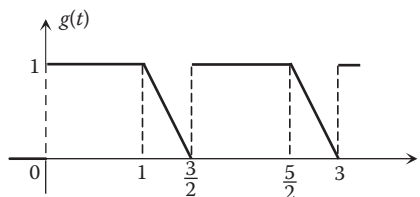


FIGURE 2.20
Problem 12.

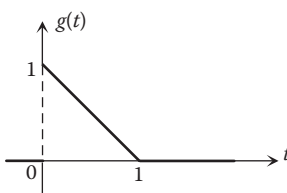


FIGURE 2.21
Problem 15.

16. a. Using partial-fraction expansion, find

$$\mathcal{L}^{-1}\left\{\frac{4s^3 + 8s^2 + 2}{s^2(4s^2 + 1)}\right\}$$

- b.  Confirm the result of (a) in MATLAB.

17. a. Find

$$\mathcal{L}^{-1}\left\{\frac{e^{-s}(4s^2 + 12s + 5)}{2s(4s^2 + 8s + 5)}\right\}$$

- b.  Confirm the result of (a) in MATLAB.

18. Consider

$$X(s) = \frac{s^2 + 3s + 1}{s(s+1)^2}$$

- a. Using the final-value theorem, if applicable, evaluate x_{ss} .
b. Confirm the result of (a) by evaluating $\lim_{t \rightarrow \infty} x(t)$.

19. Consider

$$X(s) = \frac{2(25s^2 - 30s + 113)}{s(25s^2 + 10s + 226)}$$

- a. Using the initial-value theorem, evaluate $x(0^+)$.
b. Confirm the result of (a) by evaluating $\lim_{t \rightarrow 0} x(t)$.

20. Consider

$$X(s) = \frac{0.2s + 0.3}{s(3s^2 + 2.5s + 0.5)}$$

- a. Evaluate $\dot{x}_{ss}$ by using the final-value theorem.
b. Confirm the result of (a) by evaluating $\lim_{t \rightarrow \infty} \dot{x}(t)$.



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3

Matrix Analysis

In this chapter, we discuss the fundamentals of matrix analysis, including matrix operations and properties, as well as matrix characteristics such as the rank, the determinant, and the inverse. Matrix analysis plays a particularly important role in the treatment of systems of algebraic and/or differential equations that are coupled, that is, when several unknown variables are involved in several equations in the system model. Also discussed in this chapter is the matrix eigenvalue problem, which plays a key role in the determination of a system's natural frequencies, as well as the solution process for a system of differential equations. First, we focus on algebraic systems and then extend the ideas to systems of differential equations and the matrix eigenvalue problem.

3.1 Vectors and Matrices

An n -dimensional vector $\mathbf{v}$ is an ordered set of n scalars, written as

$$\mathbf{v} = \begin{Bmatrix} v_1 \\ v_2 \\ \dots \\ v_n \end{Bmatrix}$$

where each v_i ($i = 1, 2, \dots, n$) is a component of vector $\mathbf{v}$. A matrix is a collection of numbers (real or complex), or possibly functions, arranged in a rectangular array and enclosed by square brackets. Each of the elements in a matrix is called an entry of the matrix. The horizontal and vertical levels of entries are the rows and columns of the matrix, respectively. The number of rows and columns of a matrix determines the size of that matrix. If a matrix $\mathbf{A}$ has m rows and n columns, then its size is $m \times n$. If a matrix has the same number of rows as columns, it is a square matrix. Otherwise, it is rectangular. It is customary to denote matrices by bold-faced capital letters, such as $\mathbf{A}$. The abbreviated form of an $m \times n$ matrix is

$$\mathbf{A} = [a_{ij}]_{m \times n}, \quad i = 1, 2, \dots, m, \quad j = 1, 2, \dots, n$$

where a_{ij} is known as the (i, j) entry of $\mathbf{A}$, located at the intersection of the i th row and the j th column of $\mathbf{A}$. For instance, a_{23} is the entry located at the intersection of the second row and the third column. If $\mathbf{A}$ is a square matrix ($m = n$), the elements $a_{11}, a_{22}, \dots, a_{nn}$ are the diagonal entries of $\mathbf{A}$. These diagonal elements form the main diagonal of $\mathbf{A}$. Two matrices $\mathbf{A} = [a_{ij}]$ and $\mathbf{B} = [b_{ij}]$ are equal if they have the same size and the same entries in the respective locations. If any rows or columns (or possibly both) of $\mathbf{A}$ are deleted, a submatrix of $\mathbf{A}$ is generated. The sum of $\mathbf{A} = [a_{ij}]_{m \times n}$ and $\mathbf{B} = [b_{ij}]_{m \times n}$ is $\mathbf{C} = [c_{ij}]_{m \times n} = [a_{ij} + b_{ij}]_{m \times n}$.

$$\begin{array}{c}
 \text{\textit{i}th row} \\
 \left[\begin{array}{cccccc} a_{11} & a_{12} & \dots & \dots & a_{1n} \\ \dots & \dots & \dots & \dots & \dots \\ a_{i1} & a_{i2} & \dots & \dots & a_{in} \\ \dots & \dots & \dots & \dots & \dots \\ a_{m1} & a_{m2} & \dots & \dots & a_{mn} \end{array} \right]_{m \times n}
 \end{array}
 \begin{array}{c}
 \text{\textit{j}th column} \\
 \left[\begin{array}{ccccc} b_{11} & \dots & b_{1j} & \dots & b_{1p} \\ b_{21} & \dots & b_{2j} & \dots & b_{2p} \\ \dots & \dots & \dots & \dots & \dots \\ \dots & \dots & \dots & \dots & \dots \\ b_{n1} & \dots & b_{nj} & \dots & b_{np} \end{array} \right]_{n \times p}
 \end{array}
 =
 \left[\begin{array}{cccccc} c_{11} & \dots & c_{1j} & \dots & c_{1p} \\ \dots & \dots & \dots & \dots & \dots \\ c_{i1} & \dots & c_{ij} & \dots & c_{ip} \\ \dots & \dots & \dots & \dots & \dots \\ c_{m1} & \dots & c_{mj} & \dots & c_{mp} \end{array} \right]_{m \times p}$$

FIGURE 3.1

Construction of the matrix product.

The product of a scalar k and matrix $\mathbf{A} = [a_{ij}]_{m \times n}$ is $k\mathbf{A} = [ka_{ij}]_{m \times n}$. Consider $\mathbf{A} = [a_{ij}]_{m \times n}$ and $\mathbf{B} = [b_{ij}]_{n \times p}$, so that the number of columns of $\mathbf{A}$ is equal to the number of rows of $\mathbf{B}$. Then, their product $\mathbf{C} = \mathbf{AB}$ is $m \times p$, whose entries are obtained as

$$c_{ij} = \sum_{k=1}^n a_{ik} b_{kj}, \quad i = 1, 2, \dots, m, \quad j = 1, 2, \dots, p$$

This is schematically shown in Figure 3.1. If the number of columns of $\mathbf{A}$ does not match the number of rows of $\mathbf{B}$, the product is undefined. If the product is defined, then the (i, j) entry of $\mathbf{C}$ is simply the dot (inner) product of the i th row of $\mathbf{A}$ and the j th column of $\mathbf{B}$.

Given an $m \times n$ matrix $\mathbf{A}$, its transpose, denoted by $\mathbf{A}^T$, is an $n \times m$ matrix whose first row is the first column of $\mathbf{A}$, second row is the second column of $\mathbf{A}$, and so on. Given that all matrix operations are valid,

- $(\mathbf{A} + \mathbf{B})^T = \mathbf{A}^T + \mathbf{B}^T$
- $(k\mathbf{A})^T = k\mathbf{A}^T$, $k = \text{scalar}$
- $(\mathbf{AB})^T = \mathbf{B}^T \mathbf{A}^T$

Example 3.1: Transpose

Given

$$\mathbf{A} = \begin{bmatrix} 1 & 5 \\ -2 & 4 \end{bmatrix}, \quad \mathbf{b} = \begin{bmatrix} 2 & -3 \end{bmatrix}$$

- a. Find $\mathbf{Ab}^T$.
- b.  Verify (a) in MATLAB.

Solution

- a. Since $\mathbf{b}$ is 1×2 , the size of $\mathbf{b}^T$ is 2×1 , and hence, the operation $\mathbf{Ab}^T$ is valid with size 2×1 :

$$\mathbf{Ab}^T = \begin{bmatrix} 1 & 5 \\ -2 & 4 \end{bmatrix} \begin{Bmatrix} 2 \\ -3 \end{Bmatrix} = \begin{Bmatrix} -13 \\ -16 \end{Bmatrix}$$

b. 

```
>> A = [1 5;-2 4]; b = [2 -3];
>> A*b.'
ans =
    -13
    -16
```

3.1.1 Special Matrices

A square matrix $\mathbf{A}$ is symmetric if $\mathbf{A}^T = \mathbf{A}$ and skew-symmetric if $\mathbf{A}^T = -\mathbf{A}$. A square matrix $\mathbf{A}_{n \times n} = [a_{ij}]$ is upper-triangular if $a_{ij} = 0$ for all $i > j$; that is, all entries below the main diagonal are zeros; lower-triangular if $a_{ij} = 0$ for all $i < j$; that is, all elements above the main diagonal are zeros; and diagonal if $a_{ij} = 0$ for all $i \neq j$. In the upper- and lower-triangular matrices, all the diagonal elements may be zeros. However, in a diagonal matrix, at least one diagonal entry must be non-zero. The $n \times n$ identity matrix, denoted by $\mathbf{I}$, is a diagonal matrix whose diagonal entries are equal to 1.

3.1.2 Elementary Row Operations

Matrices generally do not appear in a special form such as triangular or diagonal. However, because special matrices are so much easier to work with, a general matrix is often transformed (or reduced) into one of the mentioned special forms. This is done with the aid of elementary row operations (EROs). There are three types of EROs:

- ERO₁: Multiply a row by a non-zero constant
- ERO₂: Interchange two rows
- ERO₃: Multiply the i th row by $\alpha = \text{const} \neq 0$, add the result to the k th row, and then replace the k th row with the outcome. In this process, the i th row is called the pivot row.

It is important to note that *a matrix and its transformed form are not the same*, but depending on the transformation, certain characteristics of the original matrix may be preserved in the process. These EROs are often used to transform a matrix $\mathbf{A}$ into an upper-triangular form. The form of matrix $\mathbf{A}$ in the final stage of this process is called the row-echelon form of $\mathbf{A}$, denoted by $\text{REF}(\mathbf{A})$. We reiterate that the original matrix and any subsequent one generated by an ERO are completely different matrices. Furthermore, the REF of a matrix is not unique because various sequences of EROs may lead to an upper-triangular matrix. Although $\text{REF}(\mathbf{A})$ is not unique for a given matrix $\mathbf{A}$, important characteristics of $\mathbf{A}$ are preserved, regardless of the choice of the REF.

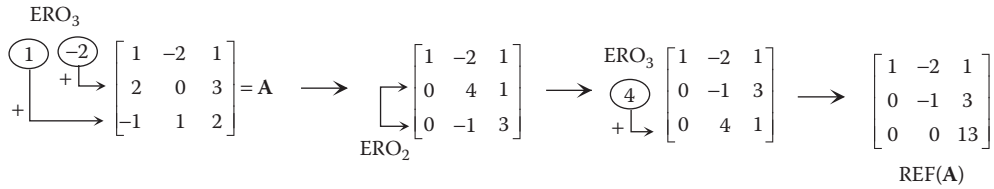
Example 3.2: Row-Echelon Forms and Elementary Row Operations

Find the REF of

$$\mathbf{A} = \begin{bmatrix} 1 & -2 & 1 \\ 2 & 0 & 3 \\ -1 & 1 & 2 \end{bmatrix}$$

Solution

The details are shown in [Figure 3.2](#). In the first segment, the first row is the pivot row and used to generate zeros below the (1,1) entry. In the second segment, the lower two

**FIGURE 3.2**

Steps in the construction of the row-echelon form.

rows are switched to replace the entry of 4 with -1 to avoid fractions. In the third segment, the new second row is the pivot row, which is used to create a zero below the $(2, 2)$ entry. The final matrix represents $\text{REF}(\mathbf{A})$. To address the nonuniqueness of the REF, note that if the two rows were not swapped in the second segment, the upper-triangular matrix in the final step may have been

$$\begin{bmatrix} 1 & -2 & 1 \\ 0 & 4 & 1 \\ 0 & 0 & 1 \end{bmatrix}$$

3.1.3 Rank of a Matrix

The rank of a matrix $\mathbf{A}$, denoted by $\text{rank}(\mathbf{A})$, is the number of non-zero rows in the REF of $\mathbf{A}$. This number is independent of the sequence of EROs used. The rank is sometimes called the row rank. If $\mathbf{A}$ is $m \times n$, then $\text{rank}(\mathbf{A})$ can at most be equal to m or n , whichever is smaller. For instance, the rank of a 3×4 matrix can at most be 3. If elementary column operations are employed to find the column echelon form, then the number of non-zero columns is the column rank of the matrix. It turns out that the row rank and the column rank of a matrix are the same, and consequently,

$$\text{rank}(\mathbf{A}) = \text{rank}(\mathbf{A}^T)$$

Example 3.3: Rank

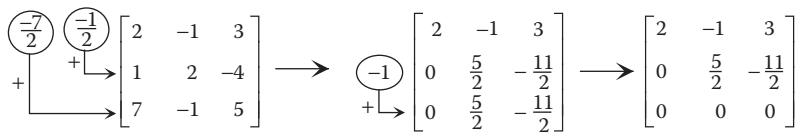
- Find the rank of the given matrix.
- Confirm the result of (a) in MATLAB.

$$\mathbf{A} = \begin{bmatrix} 2 & -1 & 3 \\ 1 & 2 & -4 \\ 7 & -1 & 5 \end{bmatrix}$$

Solution

- The details are shown in [Figure 3.3](#).
Since there are two non-zero rows in the REF of $\mathbf{A}$, we have $\text{rank}(\mathbf{A}) = 2$.
-

```
>> A = [2 -1 3; 1 2 -4; 7 -1 5];
>> rank(A)
ans =
    2
```

**FIGURE 3.3**

Construction of row-echelon form in Example 3.3.

3.1.4 Determinant of a Matrix

The determinant of a square matrix $\mathbf{A} = [a_{ij}]_{n \times n}$ is a scalar denoted by $|\mathbf{A}|$ or $\det(\mathbf{A})$. The most trivial case is $\mathbf{A} = [a_{11}]_{1 \times 1}$, for which the determinant is simply $|\mathbf{A}| = a_{11}$. For $n \geq 2$, the determinant may be calculated using any row—with preference given to the row with the most number of zeros—as

$$|\mathbf{A}| = \sum_{k=1}^n a_{ik} (-1)^{i+k} M_{ik}, \quad i = 1, 2, \dots, n \quad (3.1)$$

In Equation 3.1, M_{ik} is called the minor of a_{ik} , defined as the determinant of the $(n-1) \times (n-1)$ submatrix of $\mathbf{A}$, obtained by deleting the i th row and the k th column of $\mathbf{A}$. The quantity $(-1)^{i+k} M_{ik}$ is the cofactor of a_{ik} and is denoted by C_{ik} . In addition, note that $(-1)^{i+k}$ is responsible for whether a term is multiplied by +1 or -1. A square matrix with a non-zero determinant is called nonsingular. Otherwise, it is called singular.

Example 3.4: Determinant

- a. Calculate the determinant of

$$\mathbf{A} = \begin{bmatrix} -2 & 1 & 0 & 1 \\ 1 & 2 & -1 & 5 \\ 1 & 4 & 2 & 6 \\ 3 & -1 & 4 & 2 \end{bmatrix}$$

- b.  Repeat (a) in MATLAB.

Solution

- a. We will use the first row, since it contains one zero entry.

$$\begin{vmatrix} -2 & 1 & 0 & 1 \\ 1 & 2 & -1 & 5 \\ 1 & 4 & 2 & 6 \\ 3 & -1 & 4 & 2 \end{vmatrix} \xrightarrow{\text{Using the first row}} -2 \begin{vmatrix} 2 & -1 & 5 \\ 4 & 2 & 6 \\ -1 & 4 & 2 \end{vmatrix} - 1 \begin{vmatrix} 1 & -1 & 5 \\ 1 & 2 & 6 \\ 3 & 4 & 2 \end{vmatrix} + 1 \begin{vmatrix} 1 & 2 & -1 \\ 1 & 4 & 2 \\ 3 & -1 & 4 \end{vmatrix}$$

Each of the three 3×3 determinants is computed via Equation 3.1, resulting in

$$|\mathbf{A}| = -2(64) - (-46) - (35) = -117$$

- b. 

```
>> A = [-2 1 0 1; 1 2 -1 5; 1 4 2 6; 3 -1 4 2];
>> det(A)
ans =
    -117
```

3.1.4.1 Properties of Determinant

The following properties are associated with the matrix determinant:

- If an entire row (or column) of $\mathbf{A}$ is zero, then $|\mathbf{A}| = 0$, that is, $\mathbf{A}$ is singular.
- If any rows (or columns) of $\mathbf{A}$ are linearly dependent, then $\mathbf{A}$ is singular.
- If $\mathbf{A}$ is $n \times n$ and $\text{rank}(\mathbf{A}) < n$, then $\mathbf{A}$ is singular.
- $|\mathbf{A}^T| = |\mathbf{A}|$
- $|\mathbf{AB}| = |\mathbf{A}||\mathbf{B}|$
- The determinant of a lower-triangular, upper-triangular, or diagonal matrix is the product of the diagonal entries.
- If $k \neq 0$ is a scalar and $\mathbf{A}$ is $n \times n$, then $|k\mathbf{A}| = k^n |\mathbf{A}|$.

The effects of EROs on the matrix determinant:

- ERO₁: If a row of $\mathbf{A}$ is multiplied by a scalar $k \neq 0$, then the determinant of the resulting matrix is $k|\mathbf{A}|$.
- ERO₂: If any two rows of $\mathbf{A}$ are interchanged, then the determinant of the resulting matrix is $-|\mathbf{A}|$.
- ERO₃: Determinant is preserved under ERO₃.

3.1.4.2 Rank in Terms of Determinant

The rank of a matrix $\mathbf{A}$ is the size of the largest nonsingular submatrix of $\mathbf{A}$. In other words, $\text{rank}(\mathbf{A}) = r$ if there exists an $r \times r$ submatrix of $\mathbf{A}$ with non-zero determinant and any other $p \times p$ (with $p > r$) submatrix of $\mathbf{A}$ is singular. If $\mathbf{A}_{n \times n}$ is nonsingular, then $\text{rank}(\mathbf{A}) = n$. If $|\mathbf{A}| = 0$, then $\text{rank}(\mathbf{A}) < n$.

Example 3.5: Rank Via Determinant

Find the rank of

$$\mathbf{A} = \begin{bmatrix} 2 & 0 & 1 & 1 \\ 3 & -1 & 1 & 0 \\ 4 & 2 & 3 & 5 \end{bmatrix}$$

Solution

Since $\mathbf{A}$ is 3×4 , $\text{rank}(\mathbf{A})$ can at most be 3. Inspection of all possible 3×3 submatrices of $\mathbf{A}$ yields

$$\begin{vmatrix} 2 & 0 & 1 \\ 3 & -1 & 1 \\ 4 & 2 & 3 \end{vmatrix} = 0, \quad \begin{vmatrix} 2 & 0 & 1 \\ 3 & -1 & 0 \\ 4 & 2 & 5 \end{vmatrix} = 0, \quad \begin{vmatrix} 0 & 1 & 1 \\ -1 & 1 & 0 \\ 2 & 3 & 5 \end{vmatrix} = 0$$

This implies that $\text{rank}(\mathbf{A}) < 3$; hence, it may be either 2 or 1. Since $\begin{vmatrix} 2 & 0 \\ 3 & -1 \end{vmatrix} = -2 \neq 0$, we conclude that $\text{rank}(\mathbf{A}) = 2$.

3.1.4.3 Block Diagonal and Block Triangular Matrices

A block diagonal matrix is defined as a square matrix partitioned in such a way that its diagonal elements are square matrices, whereas all other elements are zeros (Figure 3.4a). Similarly, a block triangular matrix is defined as a square matrix partitioned such that its diagonal elements are square blocks, whereas all entries either above or below this main block diagonal are zeros (Figures 3.4b and c). It is interesting to note that many properties of these special block matrices essentially generalize those of diagonal and triangular matrices. In particular, the determinant of each of these matrices is equal to the product of the individual determinants of the blocks along the main diagonal. Consequently, a block diagonal (or triangular) matrix is singular if and only if one of the blocks along the main diagonal is singular. We also mention that the rank of a block triangular matrix is at least equal to the sum of the ranks of the individual diagonal blocks.

Example 3.6: Block Triangular Matrix

Evaluate the determinant and rank of

$$\mathbf{A} = \begin{bmatrix} 1 & 3 & 0 & 0 \\ -2 & 4 & 0 & 0 \\ 0 & 2 & 5 & 0 \\ 0 & 0 & 1 & 2 \end{bmatrix}$$

Solution

As observed in Figure 3.5, $\mathbf{A}$ is a block lower-triangular matrix. Each of the two 2×2 blocks has a determinant of 10 and a rank of 2; thus, the determinant of the matrix is $(10)(10) = 100$ and its rank is 4.

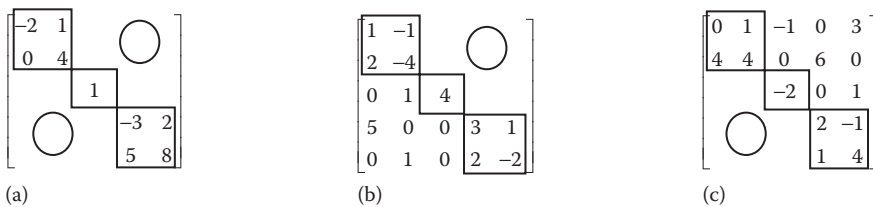


FIGURE 3.4

(a) Block diagonal, (b) block lower-triangular, and (c) block upper-triangular.

$$\mathbf{A} = \begin{bmatrix} 1 & 3 & 0 & 0 \\ -2 & 4 & 0 & 0 \\ 0 & 2 & 5 & 0 \\ 0 & 0 & 1 & 2 \end{bmatrix}$$

FIGURE 3.5

Example 3.6.

3.1.5 Inverse of a Matrix

The inverse of a square matrix $\mathbf{A}_{n \times n}$ is denoted by $\mathbf{A}^{-1}$, with the property $\mathbf{A}\mathbf{A}^{-1} = \mathbf{I} = \mathbf{A}^{-1}\mathbf{A}$, where $\mathbf{I}$ is the $n \times n$ identity matrix. The inverse of $\mathbf{A}$ exists only if $|\mathbf{A}| \neq 0$ and is obtained using the adjoint matrix of $\mathbf{A}$, denoted by $\text{adj}(\mathbf{A})$.

3.1.5.1 Adjoint Matrix

If $\mathbf{A} = [a_{ij}]_{n \times n}$, then the adjoint of $\mathbf{A}$ is defined as

$$\text{adj}(\mathbf{A}) = \begin{bmatrix} (-1)^{1+1} M_{11} & (-1)^{2+1} M_{21} & \dots & (-1)^{n+1} M_{n1} \\ (-1)^{1+2} M_{12} & (-1)^{2+2} M_{22} & \dots & (-1)^{n+2} M_{n2} \\ \dots & \dots & \dots & \dots \\ (-1)^{1+n} M_{1n} & (-1)^{2+n} M_{2n} & \dots & (-1)^{n+n} M_{nn} \end{bmatrix} = \begin{bmatrix} C_{11} & C_{21} & \dots & C_{n1} \\ C_{12} & C_{22} & \dots & C_{n2} \\ \dots & \dots & \dots & \dots \\ C_{1n} & C_{2n} & \dots & C_{nn} \end{bmatrix} \quad (3.2)$$

where M_{ij} is the minor of a_{ij} and $C_{ij} = (-1)^{i+j} M_{ij}$ is the cofactor of a_{ij} . Note that each minor M_{ij} (or cofactor C_{ij}) occupies the (j, i) position in the adjoint matrix, the opposite of what one would normally expect. Then,

$$\mathbf{A}^{-1} = \frac{1}{|\mathbf{A}|} \text{adj}(\mathbf{A}) \quad (3.3)$$

Equation 3.3 clearly indicates that $\mathbf{A}$ must be nonsingular for $\mathbf{A}^{-1}$ to exist.

Example 3.7: Inverse

- a. Find the inverse of

$$\mathbf{A} = \begin{bmatrix} 3 & 1 & -2 \\ -1 & -3 & 6 \\ -2 & 2 & 4 \end{bmatrix}$$

- b.  Verify the result of (a) in MATLAB.

Solution

- a. Noting $|\mathbf{A}| = -64 \neq 0$, the inverse exists. The first few minors and cofactors are computed as

$$C_{11} = (-1)^{1+1} M_{11} = (-1)^{1+1} \begin{vmatrix} -3 & 6 \\ 2 & 4 \end{vmatrix} = -24$$

$$C_{12} = (-1)^{1+2} M_{12} = (-1)^{1+2} \begin{vmatrix} -1 & 6 \\ -2 & 4 \end{vmatrix} = -8$$

$$C_{21} = (-1)^{2+1} M_{21} = (-1)^{2+1} \begin{vmatrix} 1 & -2 \\ 2 & 4 \end{vmatrix} = -8$$

Continuing this process, by Equation 3.3, we find

$$\text{adj}(\mathbf{A}) = \begin{bmatrix} C_{11} & C_{21} & C_{31} \\ C_{12} & C_{22} & C_{32} \\ C_{13} & C_{23} & C_{33} \end{bmatrix} = \begin{bmatrix} -24 & -8 & 0 \\ -8 & 8 & -16 \\ -8 & -8 & -8 \end{bmatrix}$$

Finally,

$$\mathbf{A}^{-1} = \frac{1}{-64} \begin{bmatrix} -24 & -8 & 0 \\ -8 & 8 & -16 \\ -8 & -8 & -8 \end{bmatrix} = \begin{bmatrix} \frac{3}{8} & \frac{1}{8} & 0 \\ \frac{1}{8} & -\frac{1}{8} & \frac{1}{4} \\ \frac{1}{8} & \frac{1}{8} & \frac{1}{8} \end{bmatrix}$$

b. 

```
>> A = [3 1 -2; -1 -3 6; -2 2 4]; inv(A)
ans =
    0.3750    0.1250         0
    0.1250   -0.1250    0.2500
    0.1250    0.1250    0.1250
```

Example 3.8: Inverse of a Symbolic Matrix

- a. Determine $(s\mathbf{I} - \mathbf{A})^{-1}$, where s is the Laplace variable, $\mathbf{I}$ is the 2×2 identity matrix, and

$$\mathbf{A} = \begin{bmatrix} 0 & 1 \\ -4 & -5 \end{bmatrix}$$

- b.  Repeat (a) in MATLAB.

Solution

- a. First

$$s\mathbf{I} - \mathbf{A} = \begin{bmatrix} s & 0 \\ 0 & s \end{bmatrix} - \begin{bmatrix} 0 & 1 \\ -4 & -5 \end{bmatrix} = \begin{bmatrix} s & -1 \\ 4 & s+5 \end{bmatrix} \Rightarrow |s\mathbf{I} - \mathbf{A}| = s^2 + 5s + 4 = (s+1)(s+4)$$

Then, by Equation 3.3,

$$(s\mathbf{I} - \mathbf{A})^{-1} = \frac{\text{adj}(s\mathbf{I} - \mathbf{A})}{|s\mathbf{I} - \mathbf{A}|} = \frac{1}{(s+1)(s+4)} \begin{bmatrix} s+5 & 1 \\ -4 & s \end{bmatrix} = \begin{bmatrix} \frac{s+5}{(s+1)(s+4)} & \frac{1}{(s+1)(s+4)} \\ \frac{-4}{(s+1)(s+4)} & \frac{s}{(s+1)(s+4)} \end{bmatrix}$$

b. 

```
>> syms s
>> A = [0 1; -4 -5]; inv(s*eye(2) - A)
ans =
[ (s+5)/(s^2+5*s+4), 1/(s^2+5*s+4) ]
[ -4/(s^2+5*s+4), s/(s^2+5*s+4) ]
```

The following properties are associated with the inverse of a matrix:

- $(\mathbf{A}^{-1})^{-1} = \mathbf{A}$
- $(\mathbf{AB})^{-1} = \mathbf{B}^{-1}\mathbf{A}^{-1}$
- $(\mathbf{A}^{-1})^p = (\mathbf{A}^p)^{-1}$, $p = \text{integer} > 0$
- $|\mathbf{A}^{-1}| = 1/|\mathbf{A}|$
- $(\mathbf{A}^{-1})^T = (\mathbf{A}^T)^{-1}$
- Inverse of a (nonsingular) symmetric matrix is symmetric.
- Inverse of a diagonal matrix is diagonal, whose entries are the reciprocals of the entries of the original matrix.

PROBLEM SET 3.1

In Problems 1 through 8, perform the indicated operations, if defined, for the following vectors and matrices.

$$\mathbf{A} = \begin{bmatrix} 0 & -2 & 1 \\ 1 & -3 & 3 \\ 2 & 1 & 2 \end{bmatrix}, \quad \mathbf{B} = \begin{bmatrix} -3 & 1 & 1 \\ 0 & 4 & 5 \end{bmatrix}, \quad \mathbf{v} = \begin{Bmatrix} 1 \\ -2 \end{Bmatrix}, \quad \mathbf{w} = \begin{Bmatrix} 1 \\ 0 \\ -1 \end{Bmatrix}$$

1. $\mathbf{w}^T \mathbf{A}$
2. $\mathbf{BAw}$
3. $(\mathbf{AB}^T)\mathbf{v}$
4. $(\mathbf{B} - \mathbf{vw}^T)\mathbf{A}$
5. $\mathbf{Aw} - \mathbf{B}^T\mathbf{v}$
6. $\mathbf{A}(\mathbf{BA} + \mathbf{vw}^T)$
7. $\mathbf{v}^T \mathbf{Bw}$
8. $(\mathbf{A}^2 + \mathbf{B}^T \mathbf{B})\mathbf{w}$

In Problems 9 through 12, find the rank by inspecting the REF of the matrix.

$$9. \mathbf{A} = \begin{bmatrix} 3 & -2 & 0 \\ -1 & 2 & 4 \\ 4 & -4 & -3 \end{bmatrix}$$

$$10. \mathbf{A} = \begin{bmatrix} 2 & -1 & 1 \\ 3 & 3 & 2 \\ 1 & -2 & 0 \end{bmatrix}$$

$$11. \mathbf{A} = \begin{bmatrix} 1 & -1 & -1 & 0 \\ 3 & -2 & 0 & 1 \\ 0 & 2 & 4 & 1 \\ 1 & 1 & 3 & 1 \end{bmatrix}$$

$$12. \mathbf{A} = \begin{bmatrix} 1 & -1 & 2 & 0 \\ 3 & 2 & 0 & 1 \\ 2 & 3 & -2 & 1 \\ 4 & 1 & 2 & 1 \end{bmatrix}$$

13. Determine the value of a such that $\text{rank}(\mathbf{A}) = 2$.

$$\mathbf{A} = \begin{bmatrix} 2 & 1 & 0 & -2 \\ 1 & 3 & -1 & 4 \\ 1 & -2 & 1 & -6 \\ 5 & a & -1 & 0 \end{bmatrix}$$

14. Given

$$\mathbf{A} = \begin{bmatrix} 0 & 1 & 0 \\ 0 & 0 & 1 \\ -1 & -2 & -1 \end{bmatrix}, \quad \mathbf{B} = \begin{Bmatrix} 0 \\ 0 \\ \frac{1}{2} \end{Bmatrix}$$

find the rank of

$$\begin{bmatrix} \mathbf{B} & \mathbf{AB} & \mathbf{A}^2\mathbf{B} \end{bmatrix}$$

In Problems 15 through 18, evaluate the determinant.

$$15. \begin{vmatrix} -1 & 2 & 3 & 1 \\ 6 & 0 & 1 & -3 \\ 0 & 1 & 5 & -1 \\ 1 & -3 & 0 & 1 \end{vmatrix}$$

$$16. \begin{vmatrix} 0 & 0 & -3 & -2 \\ 0 & b & 0 & 1 \\ a & -1 & a & 2 \\ 0 & 0 & 0 & 2a \end{vmatrix}, \quad a, b = \text{parameters}$$

$$17. \begin{vmatrix} s & -1 & 0 & 0 \\ 0 & s & 0 & -1 \\ 1 & 2 & s+1 & 0 \\ 0 & 1 & 3 & s+2 \end{vmatrix}, \quad s = \text{parameter}$$

$$18. \begin{vmatrix} 5 & -2 & 0 & 0 & 0 \\ 2 & 4 & 0 & 0 & 10 \\ 0 & 0 & -3 & 0 & 0 \\ 0 & 0 & 0 & -1 & 2 \\ 0 & 0 & 1 & 3 & 5 \end{vmatrix}$$

In Problems 19 through 24, find the inverse of the matrix.

$$19. \mathbf{A} = \begin{bmatrix} \cos \alpha & \sin \alpha \\ -\sin \alpha & \cos \alpha \end{bmatrix}, \alpha = \text{parameter}$$

$$20. \mathbf{A} = \begin{bmatrix} 1 & 0 & 2 \\ 0 & 2 & -1 \\ 0 & 0 & 3 \end{bmatrix}$$

$$21. \mathbf{A} = \begin{bmatrix} 1 & -1 & 0 \\ 3 & 0 & 0 \\ 2 & 0 & -2 \end{bmatrix}$$

$$22. \mathbf{A} = \begin{bmatrix} a+1 & 0 & 1 \\ 0 & a & 2 \\ 1 & 0 & a+2 \end{bmatrix}, a = \text{parameter}$$

$$23. \mathbf{A} = \begin{bmatrix} a & 0 & 0 \\ 0 & a+3 & 0 \\ a+1 & a+2 & 3(a+1) \end{bmatrix}, a = \text{parameter}$$

$$24. \mathbf{A} = \begin{bmatrix} -L_1 \sin \theta_1 - L_2 \sin(\theta_1 + \theta_2) & -L_2 \sin(\theta_1 + \theta_2) \\ L_1 \cos \theta_1 + L_2 \cos(\theta_1 + \theta_2) & L_2 \cos(\theta_1 + \theta_2) \end{bmatrix}, L_1, L_2, \theta_1, \theta_2 = \text{parameters}$$

25. Prove that the inverse of a (nonsingular) symmetric matrix is symmetric.

26. Given $\mathbf{A}_{n \times n}$ and scalar $k \neq 0$, show that $(k\mathbf{A})^{-1} = \frac{1}{k} \mathbf{A}^{-1}$.

3.2 Solution of Linear Systems of Equations

A linear system of n equations in n unknowns appears in the form

$$\begin{cases} a_{11}x_1 + a_{12}x_2 + \dots + a_{1n}x_n = b_1 \\ a_{21}x_1 + a_{22}x_2 + \dots + a_{2n}x_n = b_2 \\ \dots \\ a_{n1}x_1 + a_{n2}x_2 + \dots + a_{nn}x_n = b_n \end{cases} \quad (3.4)$$

where a_{ij} ($i, j = 1, 2, \dots, n$) and b_k ($k = 1, 2, \dots, n$) are the known constants, and a_{ij} 's are the coefficients. Equation 3.4 can be conveniently expressed in matrix form, as

$$\mathbf{Ax} = \mathbf{b} \quad (3.5)$$

with

$$\mathbf{A} = \begin{bmatrix} a_{11} & a_{12} & \dots & a_{1n} \\ a_{21} & a_{22} & \dots & a_{2n} \\ \dots & \dots & \dots & \dots \\ a_{n1} & a_{n2} & \dots & a_{nn} \end{bmatrix}_{n \times n}, \quad \mathbf{x} = \begin{Bmatrix} x_1 \\ x_2 \\ \dots \\ x_n \end{Bmatrix}_{n \times 1}, \quad \mathbf{b} = \begin{Bmatrix} b_1 \\ b_2 \\ \dots \\ b_n \end{Bmatrix}_{n \times 1}$$

where $\mathbf{A}$ is the coefficient matrix. A set of values for $x_1, x_2, \dots, x_n$ satisfying Equation 3.4 forms a solution of the system. The vector $\mathbf{x}$ with components $x_1, x_2, \dots, x_n$ is the solution of Equation 3.5. If $x_1 = 0 = x_2 = \dots = x_n$, the solution $\mathbf{x} = \mathbf{0}_{n \times 1}$ is called the trivial solution. The augmented matrix for Equation 3.5 is defined as

$$[\mathbf{A} | \mathbf{b}] = \left[\begin{array}{cccc|c} a_{11} & a_{12} & \dots & a_{1n} & b_1 \\ a_{21} & a_{22} & \dots & a_{2n} & b_2 \\ \dots & \dots & \dots & \dots & \dots \\ a_{n1} & a_{n2} & \dots & a_{nn} & b_n \end{array} \right]_{n \times (n+1)} \quad (3.6)$$

3.2.1 Gauss Elimination Method

Gauss elimination is a basic method to solve a linear system in the form of Equation 3.5 when n is not large. It comprises two phases: (1) Use the EROs to transform the augmented matrix $[\mathbf{A} | \mathbf{b}]$ into the upper-triangular form and (2) find the unknowns through back substitution.

Example 3.9: Gauss Elimination Method

Using Gauss elimination, solve the linear system

$$\begin{bmatrix} -1 & 1 & 3 \\ 1 & -2 & 2 \\ 2 & 1 & -4 \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \\ x_3 \end{Bmatrix} = \begin{Bmatrix} 3 \\ 9 \\ -8 \end{Bmatrix}$$

Solution

Figure 3.6 shows the steps involved in transforming the augmented matrix into the upper-triangular form, followed by back substitution.

3.2.2 Using the Inverse of the Coefficient Matrix

If the coefficient matrix $\mathbf{A}$ in Equation 3.5 is nonsingular, then the vector solution can be obtained as $\mathbf{x} = \mathbf{A}^{-1}\mathbf{b}$.

3.2.2.1 MATLAB Built-in Function "\"

The built-in function in MATLAB for solving a linear system $\mathbf{Ax}=\mathbf{b}$ is the backslash ($\backslash$), and the solution vector is obtained via $\mathbf{x}=\mathbf{A} \backslash \mathbf{b}$. It is important to note that $\mathbf{x}=\mathbf{A} \backslash \mathbf{b}$ computes the solution vector by Gauss elimination and not by $\mathbf{x}=\text{inv}(\mathbf{A}) * \mathbf{b}$.

$$\begin{array}{c}
 \textcircled{2} \quad \textcircled{1} \\
 + \quad + \\
 \downarrow \quad \downarrow \\
 \begin{bmatrix} -1 & 1 & 3 & | & 3 \\ 1 & -2 & 2 & | & 9 \\ 2 & 1 & -4 & | & -8 \end{bmatrix} \xrightarrow{\substack{R_2 + R_1 \\ R_3 + 2R_1}} \begin{bmatrix} -1 & 1 & 3 & | & 3 \\ 0 & -1 & 5 & | & 12 \\ 0 & 3 & 2 & | & -2 \end{bmatrix} \xrightarrow{\substack{R_2 \leftrightarrow R_3}} \begin{bmatrix} -1 & 1 & 3 & | & 3 \\ 0 & -1 & 5 & | & 12 \\ 0 & 0 & 17 & | & 34 \end{bmatrix}
 \end{array}$$

Upper-triangular form

$$\begin{array}{rcl}
 -x_1 + x_2 + 3x_3 = 3 & & x_1 = 1 \\
 -x_2 + 5x_3 = 12 & \Rightarrow & x_2 = -2 \\
 17x_3 = 34 & & x_3 = 2
 \end{array}$$

Back substitution

FIGURE 3.6

Gauss elimination method in Example 3.9.

Example 3.10: Solution of a Linear System

Solve the system in Example 3.9 by using

- The inverse of the coefficient matrix.
-  The "\" operator in MATLAB.

Solution

- The inverse is calculated as

$$\begin{bmatrix} -1 & 1 & 3 \\ 1 & -2 & 2 \\ 2 & 1 & -4 \end{bmatrix}^{-1} = \frac{1}{17} \begin{bmatrix} 6 & 7 & 8 \\ 8 & -2 & 5 \\ 5 & 3 & 1 \end{bmatrix}$$

Therefore,

$$\begin{Bmatrix} x_1 \\ x_2 \\ x_3 \end{Bmatrix} = \frac{1}{17} \begin{bmatrix} 6 & 7 & 8 \\ 8 & -2 & 5 \\ 5 & 3 & 1 \end{bmatrix} \begin{Bmatrix} 3 \\ 9 \\ -8 \end{Bmatrix} = \begin{Bmatrix} 1 \\ -2 \\ 2 \end{Bmatrix}$$

- 

```
>> A = [-1 1 3; 1 -2 2; 2 1 -4]; b = [3; 9; -8]; x=A\b
x =
     1
    -2
     2
```

3.2.3 Cramer's Rule

Consider the linear system in Equation 3.5. Assuming the coefficient matrix is nonsingular, each unknown is uniquely determined via

$$x_i = \frac{D_i}{D}, \quad i = 1, 2, \dots, n$$

where determinants D and D_i are formed as

$$D = \begin{vmatrix} a_{11} & a_{12} & \dots & \dots & a_{1n} \\ a_{21} & a_{22} & \dots & \dots & a_{2n} \\ \dots & \dots & \dots & \dots & \dots \\ \dots & \dots & \dots & \dots & \dots \\ a_{n1} & a_{n2} & \dots & \dots & a_{nn} \end{vmatrix}, \quad D_i = \begin{vmatrix} a_{11} & \dots & \text{\textit{ith column of D}} & \dots & a_{1n} \\ a_{21} & \dots & b_i & \dots & a_{2n} \\ \dots & \dots & \dots & \dots & \dots \\ \dots & \dots & \dots & \dots & \dots \\ a_{n1} & \dots & b_i & \dots & a_{nn} \end{vmatrix}, \quad i = 1, 2, \dots, n$$

Example 3.11: Cramer's Rule

- a. Solve the system in Example 3.9 by using Cramer's rule.
- b.  Repeat (a) in MATLAB.

Solution

- a. The determinant of the coefficient matrix is computed first,

$$D = \begin{vmatrix} -1 & 1 & 3 \\ 1 & -2 & 2 \\ 2 & 1 & -4 \end{vmatrix} = 17 \neq 0$$

Subsequently, the three determinants are calculated as

$$D_1 = \begin{vmatrix} 3 & 1 & 3 \\ 9 & -2 & 2 \\ -8 & 1 & -4 \end{vmatrix} = 17, \quad D_2 = \begin{vmatrix} -1 & 3 & 3 \\ 1 & 9 & 2 \\ 2 & -8 & -4 \end{vmatrix} = -34, \quad D_3 = \begin{vmatrix} -1 & 1 & 3 \\ 1 & -2 & 9 \\ 2 & 1 & -8 \end{vmatrix} = 34$$

Therefore, the solutions are obtained as

$$x_1 = \frac{D_1}{D} = 1, \quad x_2 = \frac{D_2}{D} = -2, \quad x_3 = \frac{D_3}{D} = 2$$

- b. 

```
>> A = [-1 1 3; 1 -2 2; 2 1 -4]; b = [3; 9; -8]; D = det(A);
>> A1 = A; A1(:,1) = b; D1 = det(A1);
>> A2 = A; A2(:,2) = b; D2 = det(A2);
>> A3 = A; A3(:,3) = b; D3 = det(A3);
>> x1 = D1/D; x2 = D2/D; x3 = D3/D; x = [x1;x2;x3]
```

```
x =
    1.0000
   -2.0000
    2.0000
```

3.2.4 Homogeneous Systems

Consider a homogeneous system $\mathbf{Ax} = \mathbf{0}$, and let $\mathbf{A}_{n \times n}$ be nonsingular. Since $\mathbf{b} = \mathbf{0}_{n \times 1}$, every determinant D_i defined in Cramer's rule contains a column of all zero entries, hence is zero. Consequently, $x_i = D_i/D = 0$ for $i = 1, 2, \dots, n$, and the only possible solution is the trivial solution $\mathbf{x} = \mathbf{0}_{n \times 1}$.

In order for $\mathbf{Ax} = \mathbf{0}$ to have a nontrivial solution, the coefficient matrix must be singular, $|\mathbf{A}| = 0$. This ensures that the equations are linearly dependent, so that there is at least one free variable that can ultimately generate infinitely many solutions.

PROBLEM SET 3.2

In Problems 1 through 6, solve the linear system $\mathbf{Ax} = \mathbf{b}$ by using Gauss elimination.

$$1. \mathbf{A} = \begin{bmatrix} 2 & 2 & -1 \\ 3 & 1 & 4 \\ -1 & 2 & 5 \end{bmatrix}, \mathbf{b} = \begin{Bmatrix} 1 \\ -2 \\ -8 \end{Bmatrix}$$

$$2. \mathbf{A} = \begin{bmatrix} -1 & 0 & 4.2 \\ 2.3 & -3 & 1.5 \\ 5 & 3.2 & 1 \end{bmatrix}, \mathbf{b} = \begin{Bmatrix} 10.4 \\ -4.6 \\ -4.8 \end{Bmatrix}$$

$$3. \mathbf{A} = \begin{bmatrix} 3 & 1.4 & -2.1 \\ 1.5 & -3.2 & 1 \\ 0.75 & 2.5 & -2 \end{bmatrix}, \mathbf{b} = \begin{Bmatrix} 4.10 \\ 11.9 \\ -4.75 \end{Bmatrix}$$

$$4. \mathbf{A} = \begin{bmatrix} 3 & 0 & -2 \\ 1 & -4 & -3 \\ 2 & 3 & 2 \end{bmatrix}, \mathbf{b} = \begin{Bmatrix} 0 \\ 1 \\ 4 \end{Bmatrix}$$

$$5. \mathbf{A} = \begin{bmatrix} 2 & -1 & 0 & 3 \\ 1 & 3 & 1 & -4 \\ 2 & 0 & 3 & 5 \\ -1 & -2 & 4 & 6 \end{bmatrix}, \mathbf{b} = \begin{Bmatrix} 10 \\ -14 \\ 9 \\ 11 \end{Bmatrix}$$

$$6. \mathbf{A} = \begin{bmatrix} -2 & 4 & 3 & 1 \\ 3 & 2 & -6 & -1 \\ 1 & -2 & -3 & 2 \\ 4 & 8 & 9 & -3 \end{bmatrix}, \mathbf{b} = \begin{Bmatrix} 7 \\ -6 \\ 1 \\ -3 \end{Bmatrix}$$

In Problems 7 through 10, solve the linear system $\mathbf{Ax} = \mathbf{b}$ by using

a. The inverse of the coefficient matrix.

b.  The "\" operator in MATLAB.

$$7. \mathbf{A} = \begin{bmatrix} 2 & 1 & -2 \\ 3 & 0 & 2 \\ -1 & 2 & -1 \end{bmatrix}, \mathbf{b} = \begin{Bmatrix} 4a \\ a \\ 0 \end{Bmatrix}, a = \text{parameter}$$

$$8. \mathbf{A} = \begin{bmatrix} a-1 & 2a+1 & -1 \\ 1 & -2a & -1 \\ -1 & 2a+1 & 3a+1 \end{bmatrix}, \mathbf{b} = \begin{Bmatrix} 5a \\ -4a \\ 7a+2 \end{Bmatrix}, a = \text{parameter}$$

$$9. \mathbf{A} = \begin{bmatrix} 2 & 2 & 0 & 0 \\ \frac{1}{2} & \frac{3}{2} & 0 & 0 \\ 0 & 0 & -\frac{1}{2} & 1 \\ 0 & 0 & 1 & -1 \end{bmatrix}, \mathbf{b} = \begin{Bmatrix} 0 \\ 1 \\ -\frac{3}{2} \\ 2 \end{Bmatrix}$$

$$10. \mathbf{A} = \begin{bmatrix} 3 & 2 & 0 & 0 \\ \frac{1}{2} & \frac{4}{3} & -1 & 0 \\ 1 & 0 & \frac{3}{2} & 1 \\ 0 & 0 & 1 & -1 \end{bmatrix}, \mathbf{b} = \begin{Bmatrix} -1 \\ -\frac{1}{6} \\ -\frac{1}{2} \\ 2 \end{Bmatrix}$$

In Problems 11 through 14,

a. Solve the linear system $\mathbf{Ax} = \mathbf{b}$ by using Cramer's rule.

b.  Repeat (a) in MATLAB.

$$11. \mathbf{A} = \begin{bmatrix} 1 & 5 & 0 \\ -1 & 2 & 1 \\ -2 & 5 & 3 \end{bmatrix}, \mathbf{b} = \begin{Bmatrix} 13 \\ 0 \\ 1 \end{Bmatrix}$$

$$12. \mathbf{A} = \begin{bmatrix} 2 & \frac{1}{2} & -1 \\ 1 & 1 & -2 \\ \frac{1}{2} & -1 & \frac{1}{2} \end{bmatrix}, \mathbf{b} = \begin{Bmatrix} 19 \\ 26 \\ -8 \end{Bmatrix}$$

$$13. \mathbf{A} = \begin{bmatrix} s^2 + s + \frac{1}{3} & -\frac{1}{3} \\ -\frac{1}{3} & s + \frac{1}{3} \end{bmatrix}, \mathbf{b} = \begin{Bmatrix} F \\ 0 \end{Bmatrix}, s, F = \text{parameters}$$

$$14. \mathbf{A} = \begin{bmatrix} 3 & 2 & 0 & 0 \\ \frac{1}{2} & -\frac{4}{3} & 1 & 0 \\ -1 & 0 & \frac{3}{2} & 1 \\ 0 & 0 & 1 & 1 \end{bmatrix}, \mathbf{b} = \begin{Bmatrix} 7 \\ -\frac{8}{3} \\ -\frac{3}{4} \\ \frac{1}{2} \end{Bmatrix}$$

In Problems 15 through 17, solve the homogeneous linear system $\mathbf{Ax} = \mathbf{0}$, where the components of $\mathbf{x}$ are $x_1, x_2, \dots, x_n$.

$$15. \mathbf{A} = \begin{bmatrix} -1 & 3 & 4 \\ 2 & -2 & 1 \\ 1 & 1 & 5 \end{bmatrix}$$

$$16. \mathbf{A} = \begin{bmatrix} 2 & 0 & 1 \\ -1 & 1 & 3 \\ 2 & -2 & 3 \end{bmatrix}$$

$$17. \mathbf{A} = \begin{bmatrix} 5 & -3 & 2 \\ 0 & 2 & 4 \\ -3 & 2 & 6 \end{bmatrix}$$

18. Determine a so that the following system has a nontrivial solution:

$$\begin{bmatrix} 2 & -1 & 0 & 0 \\ 1 & a & 0 & 0 \\ 0 & 0 & 2a & 1 \\ 0 & 0 & -1 & \frac{1}{2} \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \\ x_3 \\ x_4 \end{Bmatrix} = \begin{Bmatrix} 0 \\ 0 \\ 0 \\ 0 \end{Bmatrix}$$

3.3 Matrix Eigenvalue Problem

The matrix eigenvalue problem associated with a square matrix $\mathbf{A}_{n \times n}$ is formulated as

$$\mathbf{A}\mathbf{v} = \lambda\mathbf{v}, \quad \mathbf{v} \neq \mathbf{0}_{n \times 1}, \quad \lambda = \text{scalar} \quad (3.7)$$

A number λ , complex in general, for which Equation 3.7 has a nontrivial solution ($\mathbf{v} \neq \mathbf{0}_{n \times 1}$) is called an eigenvalue or characteristic value of matrix $\mathbf{A}$. The corresponding solution $\mathbf{v} \neq \mathbf{0}$ of Equation 3.7 is the eigenvector or characteristic vector of $\mathbf{A}$ corresponding to λ . The set of all eigenvalues of $\mathbf{A}$ forms the spectrum of $\mathbf{A}$, denoted by $\lambda(\mathbf{A})$. The largest eigenvalue of $\mathbf{A}$, in absolute value, is the spectral radius of $\mathbf{A}$.

3.3.1 Solving the Eigenvalue Problem

Rewrite Equation 3.7 as

$$\mathbf{A}\mathbf{v} - \lambda\mathbf{v} = \mathbf{0}$$

Note that every single term here is an $n \times 1$ vector. On the left-hand side, both terms contain vector $\mathbf{v}$. While in the second term λ and $\mathbf{v}$ commute, the same is not true with $\mathbf{A}$ and $\mathbf{v}$ in the first term. Therefore, we can only factor out $\mathbf{v}$ from the right to obtain

$$(\mathbf{A} - \lambda\mathbf{I})\mathbf{v} = \mathbf{0} \quad (3.8)$$

The identity matrix $\mathbf{I}$ has been inserted, so that the two terms in parentheses are size-compatible. Note that Equation 3.8 represents a homogeneous system, and, as previously mentioned, such a system has a nontrivial solution if and only if its coefficient matrix is singular. Thus, for Equation 3.8 to have a nontrivial solution ($\mathbf{v} \neq \mathbf{0}$), its coefficient matrix $\mathbf{A} - \lambda\mathbf{I}$ must be singular. In other words,

$$|\mathbf{A} - \lambda\mathbf{I}| = 0$$

This is called the characteristic equation of $\mathbf{A}$. The determinant $|\mathbf{A} - \lambda\mathbf{I}|$ is an n th-degree polynomial in λ and is known as the characteristic polynomial of $\mathbf{A}$ whose roots are the eigenvalues of $\mathbf{A}$. For each eigenvalue, we find the corresponding eigenvector by solving Equation 3.8. Since $\mathbf{A} - \lambda\mathbf{I}$ is singular, it has at least one row dependent on the other rows, which means that the REF of $\mathbf{A} - \lambda\mathbf{I}$ will have at least one zero row. Therefore, for each fixed λ , Equation 3.8 has infinitely many solutions. A basis of solutions represents all eigenvectors associated with each λ .

3.3.1.1 Eigenvalue Properties of Matrices

- Eigenvalues of upper- and lower-triangular and diagonal matrices are the entries along the main diagonal of the matrix.
- The determinant of a matrix is the product of its eigenvalues.
- All eigenvalues of a symmetric matrix are real.
- Eigenvalues of a skew-symmetric are either zero or pure imaginary.

- A matrix $\mathbf{A}$ is orthogonal if $\mathbf{A}^T = \mathbf{A}^{-1}$. All eigenvalues of an orthogonal matrix have absolute values of 1.
- The eigenvalues of block diagonal and block triangular matrices are the eigenvalues of the block matrices along the diagonal.

Example 3.12: Matrix Eigenvalue Problem

- a. Find all eigenvalues and eigenvectors of

$$\mathbf{A} = \begin{bmatrix} 1 & 3 & 1 \\ 0 & 2 & -3 \\ 0 & 0 & -2 \end{bmatrix}$$

- b.  Repeat (a) in MATLAB.

Solution

- a. Since $\mathbf{A}$ is an upper-triangular matrix, we have $\lambda(\mathbf{A}) = 1, 2, -2$. For $\lambda_1 = 1$, we solve

$$(\mathbf{A} - \lambda_1 \mathbf{I})\mathbf{v}_1 = \mathbf{0} \Rightarrow (\mathbf{A} - \mathbf{I})\mathbf{v}_1 = \mathbf{0} \Rightarrow \begin{bmatrix} 0 & 3 & 1 \\ 0 & 1 & -3 \\ 0 & 0 & -3 \end{bmatrix} \begin{Bmatrix} a \\ b \\ c \end{Bmatrix} = \begin{Bmatrix} 0 \\ 0 \\ 0 \end{Bmatrix} \xrightarrow{\text{EROs}} \begin{bmatrix} 0 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix} \begin{Bmatrix} a \\ b \\ c \end{Bmatrix} = \begin{Bmatrix} 0 \\ 0 \\ 0 \end{Bmatrix}$$

The second row gives $b=0$ and the third row gives $c=0$. The first row of zeros suggests that there is one free variable. Because $b=0$ and $c=0$, the free variable must be selected as a . Note that the free variable may be assigned any value, as long as the ensuing eigenvector is not a zero vector. For instance, in the current case, a cannot be assigned a value of zero, since b and c are already zeros. Choosing $a=1$ results in

$$\mathbf{v}_1 = \begin{Bmatrix} 1 \\ 0 \\ 0 \end{Bmatrix}$$

For $\lambda_2 = 2$, we solve

$$(\mathbf{A} - 2\mathbf{I})\mathbf{v}_2 = \mathbf{0} \Rightarrow \begin{bmatrix} -1 & 3 & 1 \\ 0 & 0 & -3 \\ 0 & 0 & -4 \end{bmatrix} \begin{Bmatrix} a \\ b \\ c \end{Bmatrix} = \begin{Bmatrix} 0 \\ 0 \\ 0 \end{Bmatrix} \xrightarrow{\text{EROs}} \begin{bmatrix} -1 & 3 & 0 \\ 0 & 0 & 1 \\ 0 & 0 & 0 \end{bmatrix} \begin{Bmatrix} a \\ b \\ c \end{Bmatrix} = \begin{Bmatrix} 0 \\ 0 \\ 0 \end{Bmatrix}$$

The second row suggests that $c=0$ and the first row suggests that $a=3b$. The last row of all zeros signifies that there is one free variable, which may be chosen as either a or b . Letting $b=1$, we have $a=3$, and

$$\mathbf{v}_2 = \begin{Bmatrix} 3 \\ 1 \\ 0 \end{Bmatrix}$$

For $\lambda_3 = -2$, solve

$$(\mathbf{A} + 2\mathbf{I})\mathbf{v}_3 = \mathbf{0} \Rightarrow \begin{bmatrix} 3 & 3 & 1 \\ 0 & 4 & -3 \\ 0 & 0 & 0 \end{bmatrix} \begin{Bmatrix} a \\ b \\ c \end{Bmatrix} = \begin{Bmatrix} 0 \\ 0 \\ 0 \end{Bmatrix}$$

The second row gives $4b = 3c$, whereas the first row yields $3a + 3b + c = 0$. The last row suggests one free variable. Expressing two of the three variables in terms of a third one, we find

$$b = \frac{3}{4}c, \quad a = -\frac{13}{12}c$$

This way, c is selected as the free variable. Letting $c = 12$, we find

$$\mathbf{v}_3 = \begin{Bmatrix} -13 \\ 9 \\ 12 \end{Bmatrix}$$

b. 

In order to find the eigenvalues and the eigenvectors, we use the command "eig" in the form of $[\mathbf{V}, \mathbf{D}] = \text{eig}(\mathbf{A})$, so that matrix $\mathbf{V}$ contains the eigenvectors (normalized as unit vectors) in its columns and diagonal matrix $\mathbf{D}$ has the eigenvalues along its main diagonal. The eigenvalue in the (1,1) entry of $\mathbf{D}$ corresponds to the eigenvector in the first column of $\mathbf{V}$, and so on.

```
>> A = [1 3 1; 0 2 -3; 0 0 -2];
>> [V,D] = eig(A)
V =
    1.0000    0.9487   -0.6549
         0    0.3162    0.4534
         0         0    0.6046

D =
     1     0     0
     0     2     0
     0     0    -2
```

As mentioned earlier, the eigenvectors are normalized to unit vectors by dividing the vector by its length (norm). For instance, eigenvector $\mathbf{v}_2$ in (a) has length $\sqrt{10}$ and

$$\frac{1}{\sqrt{10}}\mathbf{v}_2 = \frac{1}{\sqrt{10}} \begin{Bmatrix} 3 \\ 1 \\ 0 \end{Bmatrix} = \begin{Bmatrix} 0.9487 \\ 0.3162 \\ 0 \end{Bmatrix}$$

This, of course, agrees with the second column of the matrix $\mathbf{V}$, returned by "eig".

3.3.2 Algebraic Multiplicity and Geometric Multiplicity

The algebraic multiplicity (AM) of an eigenvalue is the number of times it occurs. Its geometric multiplicity (GM) is the number of linearly independent eigenvectors associated

with it. For instance, in Example 3.12, each of the three eigenvalues has the AM of 1, since each occurs only once, and a GM of 1, because there is only one independent eigenvector for each. Of course, GM of an eigenvalue λ is equal to the number of zero rows in the REF of $\mathbf{A} - \lambda\mathbf{I}$. In general, we have $\text{GM} \leq \text{AM}$. Therefore, any eigenvalue with the AM of 1 automatically has a GM of 1.

Example 3.13: Algebraic Multiplicity and Geometric Multiplicity

Find all eigenvalues and eigenvectors of the following matrix, and identify the AM and GM of each eigenvalue.

$$\mathbf{A} = \begin{bmatrix} 2 & 0 & 1 \\ 0 & 2 & 4 \\ 0 & 0 & -3 \end{bmatrix}$$

Solution

Since $\mathbf{A}$ is an upper-triangular matrix, we have $\lambda(\mathbf{A}) = 2, 2, -3$, so that $\lambda = 2$ has an AM of 2 and $\lambda = -3$ has an AM of 1. To determine the GM of each eigenvalue, the corresponding eigenvalue problem needs to be solved. For $\lambda = 2$, we solve

$$(\mathbf{A} - 2\mathbf{I})\mathbf{v} = \mathbf{0} \Rightarrow \begin{bmatrix} 0 & 0 & 1 \\ 0 & 0 & 4 \\ 0 & 0 & -5 \end{bmatrix} \begin{Bmatrix} a \\ b \\ c \end{Bmatrix} = \begin{Bmatrix} 0 \\ 0 \\ 0 \end{Bmatrix} \xRightarrow{\text{EROs}} \begin{bmatrix} 0 & 0 & 1 \\ 0 & 0 & 0 \\ 0 & 0 & 0 \end{bmatrix} \begin{Bmatrix} a \\ b \\ c \end{Bmatrix} = \begin{Bmatrix} 0 \\ 0 \\ 0 \end{Bmatrix}$$

The first row gives $c = 0$. Two zero rows indicate two free variables (a and b here), so that two linearly independent eigenvectors can be obtained. This implies that $\lambda = 2$ has a GM of 2. To find the first independent eigenvector, let $a = 1$, $b = 0$, and to find the second one, let $a = 0$, $b = 1$. Therefore, the two linearly independent eigenvectors associated with $\lambda = 2$ are

$$\left\{ \begin{Bmatrix} 1 \\ 0 \\ 0 \end{Bmatrix}, \begin{Bmatrix} 0 \\ 1 \\ 0 \end{Bmatrix} \right\}$$

For $\lambda = -3$, we solve

$$(\mathbf{A} + 3\mathbf{I})\mathbf{v} = \mathbf{0} \Rightarrow \begin{bmatrix} 5 & 0 & 1 \\ 0 & 5 & 4 \\ 0 & 0 & 0 \end{bmatrix} \begin{Bmatrix} a \\ b \\ c \end{Bmatrix} = \begin{Bmatrix} 0 \\ 0 \\ 0 \end{Bmatrix}$$

As expected, there is only one free variable, and the GM is 1. Proceeding as usual, the only independent eigenvector associated with $\lambda = -3$ is obtained as

$$\left\{ \begin{Bmatrix} 1 \\ 4 \\ -5 \end{Bmatrix} \right\}$$

3.3.2.1 Generalized Eigenvectors

Suppose AM of a certain eigenvalue λ is m , so that m corresponding eigenvectors are expected for λ . However, if the GM of λ is $k < m$, then only k linearly independent eigenvectors will be generated for λ . This means that there are $m - k$ missing eigenvectors for λ . These missing eigenvectors are called generalized eigenvectors and can be found by using a systematic approach (see Reference [6]). Any matrix with a generalized eigenvector is called *defective*. The following example illustrates how defective matrices can be handled in MATLAB.

Example 3.14: Generalized Eigenvectors

Find the eigenvalues and eigenvectors of

$$\mathbf{A} = \begin{bmatrix} 6 & 1 & -2 \\ 2 & 6 & 2 \\ 0 & -1 & 8 \end{bmatrix}$$

Solution

The eigenvalues of $\mathbf{A}$ are calculated as $\lambda(\mathbf{A}) = 8, 6, 6$, so that $\lambda_1 = 8$ has an AM of 1 and $\lambda_2 = 6$ has an AM of 2. For $\lambda_1 = 8$, we solve

$$[\mathbf{A} - 8\mathbf{I}]\mathbf{v}_1 = \mathbf{0} \Rightarrow \left[\begin{array}{ccc|c} -2 & 1 & -2 & 0 \\ 2 & -2 & 2 & 0 \\ 0 & -1 & 0 & 0 \end{array} \right] \xrightarrow{\text{EROs}} \left[\begin{array}{ccc|c} 1 & 0 & 1 & 0 \\ 0 & 1 & 0 & 0 \\ 0 & 0 & 0 & 0 \end{array} \right] \Rightarrow \mathbf{v}_1 = \begin{Bmatrix} -1 \\ 0 \\ 1 \end{Bmatrix}$$

Since there is one independent eigenvector, $\lambda_1 = 8$ has a GM of 1, as expected. For $\lambda_2 = 6$, we solve

$$[\mathbf{A} - 6\mathbf{I}]\mathbf{v}_2 = \mathbf{0} \Rightarrow \left[\begin{array}{ccc|c} 0 & 1 & -2 & 0 \\ 2 & 0 & 2 & 0 \\ 0 & -1 & 2 & 0 \end{array} \right] \xrightarrow{\text{EROs}} \left[\begin{array}{ccc|c} 0 & 1 & -2 & 0 \\ 1 & 0 & 1 & 0 \\ 0 & 0 & 0 & 0 \end{array} \right] \Rightarrow \mathbf{v}_2 = \begin{Bmatrix} -1 \\ 2 \\ 1 \end{Bmatrix}$$

Because there is one free variable, only one independent eigenvector is generated and $\lambda_1 = 6$ has a GM of 1. We now make the following observation. Matrix $\mathbf{A}$ has three eigenvalues; thus, three eigenvectors are expected. One of the eigenvalues ($\lambda_1 = 8$) occurs once and expectedly has one corresponding eigenvector. The other eigenvalue ($\lambda_2 = 6$) occurs twice, so that two eigenvectors are warranted. Since only one was found, there exists one generalized eigenvector.

3.3.2.2 Generalized Eigenvectors in MATLAB

To find the eigenvalues and eigenvectors of the current matrix $\mathbf{A}$ in MATLAB, we would ordinarily use the "eig" function, as follows:

```
>> A = [6 1 -2; 2 6 2; 0 -1 8]; [V,D] = eig(A)
V =
-0.7071    0.4082   -0.4082
 0.0000   -0.8165    0.8165
 0.7071   -0.4082    0.4082
```

```
D =
      8.0000      0      0
      0      6.0000      0
      0      0      6.0000
```

However, we notice that the second and third columns of V represent the same eigenvector. This indicates that there exists a generalized eigenvector. In order to find this missing eigenvector, we switch from "eig" to "jordan", which is specifically designed for this purpose.

```
>> [V,J] = jordan(A)
V =
      0.5000     -1.0000      0.5000
           0      2.0000           0
     -0.5000      1.0000      0.5000

J =
      8      0      0
      0      6      1
      0      0      6
```

Note that matrix J is no longer diagonal and is known as a Jordan matrix. The eigenvalues of A are 8, 6, 6. The first column of V contains the eigenvector for $\lambda = 8$. The next two columns correspond to $\lambda = 6$, with the last one being a generalized eigenvector.

3.3.2.3 Similarity Transformations

Two matrices $A_{n \times n}$ and $B_{n \times n}$ are similar if there exists a nonsingular matrix $S_{n \times n}$ such that

$$B = S^{-1}AS \quad (3.9)$$

We say that B is obtained from A through a similarity transformation. Eigenvalues of a matrix are preserved under similarity transformation; that is, $\lambda(A) = \lambda(B)$. Similarity transformations are often utilized to transform a matrix into a diagonal matrix, with eigenvectors playing a key role in that process.

3.3.2.4 Matrix Diagonalization

Suppose $A_{n \times n}$ has eigenvalues $\lambda_1, \lambda_2, \dots, \lambda_n$ and corresponding linearly independent eigenvectors $v_1, v_2, \dots, v_n$, with no generalized eigenvectors. Form the modal matrix $V_{n \times n} = [v_1 \ v_2 \ \dots \ v_n]$, which is guaranteed to be nonsingular, because its columns are linearly independent. Then,

$$V^{-1}AV = \begin{bmatrix} \lambda_1 & 0 & \dots & 0 \\ 0 & \lambda_2 & 0 & \dots \\ \dots & 0 & \dots & 0 \\ 0 & \dots & 0 & \lambda_n \end{bmatrix} = D \quad (3.10)$$

Matrix A has clearly been transformed into a diagonal matrix D by a similarity transformation.

3.3.2.5 Defective Matrices

Suppose $\mathbf{A}_{n \times n}$ has eigenvalues $\lambda_1, \lambda_2, \dots, \lambda_n$ and linearly independent eigenvectors $\mathbf{v}_1, \mathbf{v}_2, \dots, \mathbf{v}_n$, including at least one generalized eigenvector. Again, the modal matrix $\mathbf{V}_{n \times n} = [\mathbf{v}_1 \ \mathbf{v}_2 \ \dots \ \mathbf{v}_n]$ is guaranteed to be nonsingular, because its columns are linearly independent, and

$$\mathbf{V}^{-1}\mathbf{A}\mathbf{V} = \mathbf{J} \quad (3.11)$$

where $\mathbf{J}$ is not diagonal and is called a Jordan matrix (see Example 3.14).

Example 3.15: Diagonalization

In Example 3.12, we found that $\lambda(\mathbf{A}) = 1, 2, -2$ and there are no generalized eigenvectors. The modal matrix then transforms $\mathbf{A}$ into the diagonal matrix $\mathbf{D}$, as in Equation 3.10.

```
>> A = [1 3 1; 0 2 -3; 0 0 -2]; [V,D] = eig(A);
>> D = V\A*V      % Perform D = V^-1AV
D =
    1.0000    -0.0000     0.0000
         0     2.0000    -0.0000
         0         0    -2.0000
```

Example 3.16: Jordan Matrix

In Example 3.14, we found that $\lambda(\mathbf{A}) = 8, 6, 6$ and there is one generalized eigenvector. The modal matrix transforms $\mathbf{A}$ into a Jordan matrix $\mathbf{J}$, as in Equation 3.11.

```
>> A = [6 1 -2; 2 6 2; 0 -1 8]; [V,J] = jordan(A);
>> V\A*V      % Perform J = V^-1AV
ans =
     8     0     0
     0     6     1
     0     0     6
```

This, of course, agrees with the Jordan matrix $\mathbf{J}$ returned by `jordan(A)`.

PROBLEM SET 3.3

In Problems 1 through 8,

- Find the eigenvalues and eigenvectors of the matrix.
-  Confirm the results of (a) in MATLAB.

1. $\mathbf{A} = \begin{bmatrix} 2 & 4 \\ 0 & -3 \end{bmatrix}$

2. $\mathbf{A} = \begin{bmatrix} 5 & -2 \\ 2 & 0 \end{bmatrix}$

3. $\mathbf{A} = \begin{bmatrix} 0 & 3 \\ 3 & 0 \end{bmatrix}$

$$4. \mathbf{A} = \begin{bmatrix} 1 & 1 & 0 \\ 3 & 3 & 0 \\ 0 & 0 & -2 \end{bmatrix}$$

$$5. \mathbf{A} = \begin{bmatrix} 3 & -4 & -2 \\ -1 & 4 & 1 \\ 2 & -6 & -1 \end{bmatrix}$$

$$6. \mathbf{A} = \begin{bmatrix} 1 & -2 & 4 \\ -1 & 2 & 2 \\ -5 & 10 & -2 \end{bmatrix}$$

$$7. \mathbf{A} = \begin{bmatrix} 0 & 2 & -2 & 0 \\ -1 & -5 & 2 & -2 \\ 0 & -3 & 0 & -3 \\ 1 & 2 & -2 & -1 \end{bmatrix}$$

In Problems 8 through 15, find the eigenvalues, eigenvectors, AM, and GM of each eigenvalue and decide whether the matrix is defective or not. Then, transform the matrix into either a diagonal or a Jordan matrix, whichever is applicable.

$$8. \mathbf{A} = \begin{bmatrix} 2 & -1 \\ 6 & 7 \end{bmatrix}$$

$$9. \mathbf{A} = \begin{bmatrix} 0 & -1 \\ 1 & 2 \end{bmatrix}$$

$$10. \mathbf{A} = \begin{bmatrix} 3 & 0 & 0 \\ 0 & 3 & 1 \\ 5 & 0 & 0 \end{bmatrix}$$

$$11. \mathbf{A} = \begin{bmatrix} 4 & 1 & 0 \\ 0 & 4 & 1 \\ -1 & -1 & 3 \end{bmatrix}$$

$$12. \mathbf{A} = \begin{bmatrix} -2 & 3 & 1 \\ -1 & 2 & 1 \\ -1 & 1 & 0 \end{bmatrix}$$

$$13. \mathbf{A} = \begin{bmatrix} 2 & 2 & 1 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix}$$

$$14. \mathbf{A} = \begin{bmatrix} 1 & 0 & 0 \\ 0 & -2 & -1 \\ 0 & 1 & -2 \end{bmatrix}$$

$$15. \mathbf{A} = \begin{bmatrix} 5 & 0 & 0 \\ 2 & 5 & 1 \\ 0 & 0 & 6 \end{bmatrix}$$

16. Prove that if $\mathbf{A}_{2 \times 2}$ has a repeated eigenvalue λ , then $\mathbf{A}$ must be defective.
17. Prove that a singular matrix must have at least one zero eigenvalue.
18. Show that if the eigenvalues of $\mathbf{A}$ are $\lambda_1, \lambda_2, \dots, \lambda_n$, then the eigenvalues of $\mathbf{A} - \alpha \mathbf{I}$ are $\lambda_1 - \alpha, \lambda_2 - \alpha, \dots, \lambda_n - \alpha$.
19. Show that the eigenvalues of $\mathbf{A}$ and $\mathbf{A}^T$ are the same.

3.4 Summary

A matrix is a collection of elements arranged in a rectangular array and enclosed by square brackets. The size of a matrix $\mathbf{A}$ is $m \times n$ if it has m rows and n columns. The abbreviated form of an $m \times n$ matrix is

$$\mathbf{A} = [a_{ij}]_{m \times n}, \quad i = 1, 2, \dots, m, \quad j = 1, 2, \dots, n$$

Matrix addition is performed entry-wise. If k is scalar, then $k\mathbf{A} = [ka_{ij}]_{m \times n}$. If $\mathbf{A} = [a_{ij}]_{m \times n}$ and $\mathbf{B} = [b_{ij}]_{n \times p}$, then $\mathbf{C} = \mathbf{AB}$ is an $m \times p$ matrix, whose entries are obtained as

$$c_{ij} = \sum_{k=1}^n a_{ik} b_{kj}, \quad i = 1, 2, \dots, m, \quad j = 1, 2, \dots, p$$

The transpose of $\mathbf{A}_{m \times n}$, denoted by $\mathbf{A}^T$, is an $n \times m$ matrix, whose rows are the columns of $\mathbf{A}$. A matrix $\mathbf{A} = [a_{ij}]_{m \times n}$ is symmetric if $\mathbf{A}^T = \mathbf{A}$, upper-triangular if $a_{ij} = 0$ for all $i > j$, lower-triangular if $a_{ij} = 0$ for all $i < j$, and diagonal if $a_{ij} = 0$ for all $i \neq j$. Matrix transformations may be achieved by using EROs:

- ERO₁: Multiply a row by a non-zero constant
- ERO₂: Interchange two rows
- ERO₃: Multiply the i th row by $\alpha = \text{const} \neq 0$, add the result to the k th row, and then replace the k th row with the outcome. In this process, the i th row is called the pivot row.

The rank of a matrix is the number of non-zero rows in the REF of that matrix. The determinant of $\mathbf{A}_{n \times n}$ is a real scalar, calculated as

$$|\mathbf{A}| = \sum_{k=1}^n a_{ik} (-1)^{i+k} M_{ik}, \quad i = 1, 2, \dots, n$$

where M_{ik} is the minor of a_{ik} and $(-1)^{i+k} M_{ik}$ is the cofactor of a_{ik} . The inverse of matrix $\mathbf{A}$ is obtained as

$$\mathbf{A}^{-1} = \frac{1}{|\mathbf{A}|} \text{adj}(\mathbf{A})$$

where $\text{adj}(\mathbf{A})$ is the adjoint matrix of $\mathbf{A}$ and is defined as

$$\text{adj}(\mathbf{A}) = \begin{bmatrix} (-1)^{1+1} M_{11} & (-1)^{2+1} M_{21} & \dots & (-1)^{n+1} M_{n1} \\ (-1)^{1+2} M_{12} & (-1)^{2+2} M_{22} & \dots & (-1)^{n+2} M_{n2} \\ \dots & \dots & \dots & \dots \\ (-1)^{1+n} M_{1n} & (-1)^{2+n} M_{2n} & \dots & (-1)^{n+n} M_{nn} \end{bmatrix} = \begin{bmatrix} C_{11} & C_{21} & \dots & C_{n1} \\ C_{12} & C_{22} & \dots & C_{n2} \\ \dots & \dots & \dots & \dots \\ C_{1n} & C_{2n} & \dots & C_{nn} \end{bmatrix}$$

The eigenvalue problem associated with a matrix $\mathbf{A}$ is described by

$$\mathbf{A}_{n \times n} \mathbf{v}_{n \times 1} = \lambda_{1 \times 1} \mathbf{v}_{n \times 1}, \quad \mathbf{v}_{n \times 1} \neq \mathbf{0}_{n \times 1}$$

where λ is an eigenvalue of $\mathbf{A}$ and $\mathbf{v}$ is the corresponding eigenvector. The eigenvalues are the roots of the characteristic equation $|\mathbf{A} - \lambda \mathbf{I}| = 0$. The AM of an eigenvalue is the number of times it occurs. The GM of an eigenvalue is the number of linearly independent eigenvectors associated with it. In general, we have $\text{GM} \leq \text{AM}$. If $\text{GM} < \text{AM}$ for an eigenvalue, then the matrix has at least one generalized eigenvector. A matrix with at least one generalized eigenvector is called defective. Any nondefective matrix $\mathbf{A}$ can be diagonalized via a similarity transformation $\mathbf{V}^{-1} \mathbf{A} \mathbf{V} = \mathbf{D}$, where $\mathbf{V}_{n \times n} = [\mathbf{v}_1 \quad \mathbf{v}_2 \quad \dots \quad \mathbf{v}_n]$ is the modal matrix of $\mathbf{A}$, whose columns are the (linearly independent) eigenvectors of $\mathbf{A}$ and $\mathbf{D}$ is a diagonal matrix composed of the eigenvalues of $\mathbf{A}$. Defective matrices are transformed into a Jordan matrix $\mathbf{J}$.

Review Problems

1. Find the value of a such that the following two matrices have the same determinant.

$$\begin{bmatrix} 1 & 3 & 0 \\ -2 & a & 0 \\ 0 & 0 & 4 \end{bmatrix}, \quad \begin{bmatrix} 4 & 0 & 0 \\ 0 & -3 & 2a \\ 0 & \frac{1}{2} & -1 \end{bmatrix}$$

2. Determine whether the following vectors are linearly independent.

$$\left\{ \begin{bmatrix} 3 \\ -1 \\ 4 \end{bmatrix}, \begin{bmatrix} 5 \\ 4 \\ 2 \end{bmatrix}, \begin{bmatrix} 2 \\ 5 \\ -2 \end{bmatrix} \right\}$$

3. Prove that the product of two symmetric matrices is not necessarily symmetric.
4. If $\mathbf{A}$ is $m \times m$ and symmetric, and $\mathbf{B}$ is a general $m \times n$ matrix, show that $\mathbf{B}^T \mathbf{A} \mathbf{B}$ is $n \times n$ and symmetric.
5. Given

$$\mathbf{A} = \begin{bmatrix} 0 & 1 & 0 \\ 0 & 0 & 1 \\ -2 & -1 & -2 \end{bmatrix}, \quad \mathbf{C} = \begin{bmatrix} 2 & 0 & \frac{1}{3} \end{bmatrix}$$

find the rank of

$$\begin{bmatrix} \mathbf{C}^T & \mathbf{A}^T \mathbf{C}^T & (\mathbf{A}^T)^2 \mathbf{C}^T \end{bmatrix}$$

6. Determine a such that $\text{rank}(\mathbf{A}) = 3$, where

$$\mathbf{A} = \begin{bmatrix} 3 & 7 & -2 & 1 \\ -1 & 4 & -3 & 2 \\ 1 & a-4 & 5 & 3 \\ 7 & 12 & 1 & 5 \end{bmatrix}$$

7. Find the inverse of the rotation matrix

$$\mathbf{R} = \begin{bmatrix} \cos \theta & \sin \theta & 0 \\ -\sin \theta & \cos \theta & 0 \\ 0 & 0 & 1 \end{bmatrix}$$

8. Find the value(s) of a for which the following system only has a trivial solution.

$$\begin{bmatrix} \frac{2}{3} & 1 & -2 \\ a & -2 & 1 \\ 1 & 1 & -5 \end{bmatrix} \begin{bmatrix} x_1 \\ x_2 \\ x_3 \end{bmatrix} = \begin{bmatrix} 0 \\ 0 \\ 0 \end{bmatrix}$$

9. Solve the following linear system by using Cramer's rule; a is a parameter.

$$\mathbf{A} \mathbf{x} = \mathbf{b}, \quad \mathbf{A} = \begin{bmatrix} 3 & 0 & -1 & 2 \\ 2 & 2 & 0 & -2 \\ -1 & 3 & 1 & -1 \\ 2 & -1 & 2 & 3 \end{bmatrix}, \quad \mathbf{x} = \begin{bmatrix} x_1 \\ x_2 \\ x_3 \\ x_4 \end{bmatrix}, \quad \mathbf{b} = \begin{bmatrix} 2a+4 \\ 2 \\ 2a-2 \\ 2a \end{bmatrix}$$

10. Solve the linear system in Problem 9 by using the inverse of the coefficient matrix.
11. Show that any matrix with distinct eigenvalues is nondefective.

12. a. Find all eigenvalues and eigenvectors of

$$\mathbf{A} = \begin{bmatrix} -3 & -2 & 0 \\ 0 & -1 & 0 \\ 1 & -1 & -2 \end{bmatrix}$$

- b.  Confirm the results of (a) in MATLAB.

13. Find the eigenvalues and the AM and GM of each, and decide whether the matrix is defective.

$$\mathbf{A} = \begin{bmatrix} 1 & 0 & 0 & 0 \\ -2 & 3 & 0 & 0 \\ 1 & 0 & -1 & 1 \\ 0 & 2 & 0 & -1 \end{bmatrix}$$

14. Prove that the eigenvalues of a matrix are preserved under a similarity transformation; that is, if $\mathbf{S}^{-1}\mathbf{A}\mathbf{S} = \mathbf{B}$, then the eigenvalues of $\mathbf{A}$ and $\mathbf{B}$ are the same.
15.  Find the modal matrix, and use it to transform $\mathbf{A}$ into a diagonal or Jordan matrix:

$$\mathbf{A} = \begin{bmatrix} -1 & 4 & -1 & 2 \\ 0 & 1 & 1 & 4 \\ 0 & 0 & -1 & 0 \\ 0 & 0 & 2 & -5 \end{bmatrix}$$

16. Show that the rotation matrix $\mathbf{R}$ of Problem 7 is orthogonal. Then, verify that each of its eigenvalues has an absolute value of 1.
17. Prove that if $k = \text{const}$ and matrix $\mathbf{A}$ has eigenvalues $\lambda_1, \dots, \lambda_n$, with corresponding eigenvectors $\mathbf{v}_1, \dots, \mathbf{v}_n$, then the eigenvalues of $k\mathbf{A}$ are $k\lambda_1, \dots, k\lambda_n$, with eigenvectors $\mathbf{v}_1, \dots, \mathbf{v}_n$.

In Problems 18 through 20, decide whether the statement is true or false.

18. If $n \geq 3$, then every $n \times n$ matrix with any repeated eigenvalues is defective.
19. The rank of a singular 4×4 matrix is 3.
20. If all eigenvalues of a matrix are real, then the matrix is symmetric.



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4

System Model Representation

This chapter presents several commonly used forms in the representation of mathematical models of dynamic systems. The derivation of the models will be introduced and thoroughly discussed in [Chapters 5](#) through [7](#). The main topics covered in this chapter include the configuration form, state-space form, input–output (I/O) equation, transfer function, and block diagram representation. Also included is how any one of these forms may be obtained from any other form. Various methods of linearization of non-linear systems, including linearization in Simulink®, are covered in the final section of this chapter.

4.1 Configuration Form

A set of coordinates that completely describes the motion of a system is known as a set of generalized coordinates. This set is not unique, so that more than one set of coordinates can be chosen for the same system. However, the number of coordinates remains the same, regardless of the set selected for a specific system. If there are n generalized coordinates, they are usually denoted by $q_1, q_2, \dots, q_n$. Suppose a dynamic system model is described by

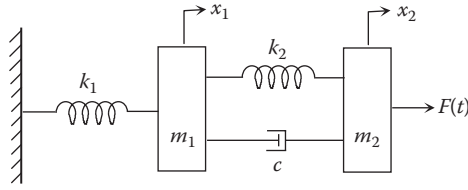
$$\begin{cases} \ddot{q}_1 = f_1(q_1, q_2, \dots, q_n, \dot{q}_1, \dot{q}_2, \dots, \dot{q}_n, t) \\ \ddot{q}_2 = f_2(q_1, q_2, \dots, q_n, \dot{q}_1, \dot{q}_2, \dots, \dot{q}_n, t) \\ \dots \\ \ddot{q}_n = f_n(q_1, q_2, \dots, q_n, \dot{q}_1, \dot{q}_2, \dots, \dot{q}_n, t) \end{cases} \quad (4.1)$$

where $\dot{q}_1, \dot{q}_2, \dots, \dot{q}_n$ are the generalized velocities and $f_1, f_2, \dots, f_n$ known as the generalized forces, are algebraic functions of q_i and $\dot{q}_i$ ($i = 1, 2, \dots, n$) and time t . Assuming initial time is $t = 0$, Equation 4.1 subjected to initial generalized coordinates $q_1(0), q_2(0), \dots, q_n(0)$ and initial generalized velocities $\dot{q}_1(0), \dot{q}_2(0), \dots, \dot{q}_n(0)$ is called the configuration form.

Example 4.1: Configuration Form

The mechanical system shown in [Figure 4.1](#) consists of blocks m_1 and m_2 , linear springs with stiffness coefficients k_1 and k_2 , a linear damper with coefficient of viscous damping c , and force $F(t)$ applied to block m_2 . The equations of motion are derived as ([Chapter 5](#))

$$\begin{cases} m_1 \ddot{x}_1 + k_1 x_1 - k_2 (x_2 - x_1) - c(\dot{x}_2 - \dot{x}_1) = 0 \\ m_2 \ddot{x}_2 + k_2 (x_2 - x_1) + c(\dot{x}_2 - \dot{x}_1) = F(t) \end{cases}$$

**FIGURE 4.1**

Mechanical system in translation.

where x_1 and x_2 are the displacements of the blocks, and $\dot{x}_1$ and $\dot{x}_2$ are their respective velocities. The system is subjected to initial displacements $x_1(0) = x_{10}$ and $x_2(0) = x_{20}$ and initial velocities $\dot{x}_1(0) = \dot{x}_{10}$ and $\dot{x}_2(0) = \dot{x}_{20}$. Obtain the configuration form.

Solution

The equations of motion are first rewritten as

$$\begin{cases} \ddot{x}_1 = \frac{1}{m_1} [-(k_1 + k_2)x_1 + k_2x_2 - c\dot{x}_1 + c\dot{x}_2] \\ \ddot{x}_2 = \frac{1}{m_2} [-k_2x_2 + k_2x_1 + c\dot{x}_1 - c\dot{x}_2 + F(t)] \end{cases}$$

Comparison with Equation 4.1 reveals that $n = 2$, so that there are two generalized coordinates: $q_1 = x_1$ and $q_2 = x_2$. The previous set of equations is then expressed as

$$\begin{aligned} \ddot{q}_1 &= \frac{1}{m_1} [-(k_1 + k_2)q_1 + k_2q_2 - c\dot{q}_1 + c\dot{q}_2] \\ \ddot{q}_2 &= \frac{1}{m_2} [-k_2q_2 + k_2q_1 + c\dot{q}_1 - c\dot{q}_2 + F(t)] \end{aligned}$$

where f_1 and f_2 are the generalized forces. These, together with the four initial conditions,

$$\begin{aligned} q_1(0) &= x_{10}, \quad \dot{q}_1(0) = \dot{x}_{10} \\ q_2(0) &= x_{20}, \quad \dot{q}_2(0) = \dot{x}_{20} \end{aligned}$$

establish the configuration form.

4.1.1 Second-Order Matrix Form

Mathematical models of dynamic systems that are governed by n -dimensional systems of second-order differential equations can conveniently be expressed as

$$\mathbf{M}\ddot{\mathbf{x}} + \mathbf{C}\dot{\mathbf{x}} + \mathbf{K}\mathbf{x} = \mathbf{f} \quad (4.2)$$

which is commonly known as the standard, second-order matrix form, where

- $\mathbf{x}_{n \times 1}$ = configuration vector and $\mathbf{f}_{n \times 1}$ = vector of external forces
- $\mathbf{M}_{n \times n}$ = mass matrix, $\mathbf{C}_{n \times n}$ = damping matrix, and $\mathbf{K}_{n \times n}$ = stiffness matrix

Example 4.2: Second-Order Matrix Form

Express the equations of motion of the system in Example 4.1 in second-order matrix form.

Solution

The equations of motion can be expressed in the form of Equation 4.2, as

$$\begin{bmatrix} m_1 & 0 \\ 0 & m_2 \end{bmatrix} \begin{Bmatrix} \ddot{x}_1 \\ \ddot{x}_2 \end{Bmatrix} + \begin{bmatrix} c & -c \\ -c & c \end{bmatrix} \begin{Bmatrix} \dot{x}_1 \\ \dot{x}_2 \end{Bmatrix} + \begin{bmatrix} k_1 + k_2 & -k_2 \\ -k_2 & k_2 \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix} = \begin{Bmatrix} 0 \\ F(t) \end{Bmatrix}$$

where the pertinent vectors and matrices are properly identified as

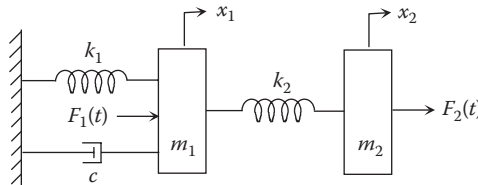
$$\mathbf{M} = \begin{bmatrix} m_1 & 0 \\ 0 & m_2 \end{bmatrix}, \quad \mathbf{C} = \begin{bmatrix} c & -c \\ -c & c \end{bmatrix}, \quad \mathbf{K} = \begin{bmatrix} k_1 + k_2 & -k_2 \\ -k_2 & k_2 \end{bmatrix}, \quad \mathbf{x} = \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix}, \quad \mathbf{f} = \begin{Bmatrix} 0 \\ F(t) \end{Bmatrix}$$

PROBLEM SET 4.1

In Problems 1 through 8, assuming general initial conditions, express the system model in

- Configuration form.
- Standard, second-order matrix form.

- $$\begin{cases} \frac{1}{2}\ddot{x}_1 + \dot{x}_1 + \frac{1}{3}(x_1 - x_2) = 0 \\ \ddot{x}_2 + \frac{1}{2}\dot{x}_2 - \frac{1}{3}(x_1 - x_2) = e^{-2t/3} \end{cases}$$
- $$\begin{cases} \frac{1}{5}\ddot{x}_1 + \dot{x}_1 + 2(x_1 - x_2) = e^{-2t} \\ \frac{3}{5}\ddot{x}_2 - 2(x_1 - x_2) = 0 \end{cases}$$
- $$\begin{cases} m_1\ddot{x}_1 + k_1x_1 + c\dot{x}_1 - k_2(x_2 - x_1) = F_1(t) \\ m_2\ddot{x}_2 + k_2(x_2 - x_1) = F_2(t) \end{cases}; \text{ mechanical system in Figure 4.2}$$
- $$\begin{cases} m_1\ddot{x}_1 + c_1\dot{x}_1 + k_1x_1 - k_2(x_2 - x_1) - c_2(\dot{x}_2 - \dot{x}_1) = F(t) \\ m_2\ddot{x}_2 + k_2(x_2 - x_1) + c_2(\dot{x}_2 - \dot{x}_1) = 0 \end{cases}; \text{ mechanical system in Figure 4.3}$$
- $$\begin{cases} \ddot{\theta}_1 + \dot{\theta}_1 + \frac{3}{2}(\theta_1 - \theta_2) = e^{-t/2} \sin t \\ \ddot{\theta}_2 + \dot{\theta}_2 - \frac{3}{2}(\theta_1 - \theta_2) = 0 \end{cases}$$
- $$\begin{cases} m\ddot{x}_1 + kx_1 - \frac{1}{2}k(x_2 - x_1) - 2c(\dot{x}_2 - \dot{x}_1) = F_1(t) \\ m\ddot{x}_2 - \frac{2}{3}k(x_3 - x_2) + \frac{1}{2}k(x_2 - x_1) + 2c(\dot{x}_2 - \dot{x}_1) = 0 \\ 2m\ddot{x}_3 + kx_3 + \frac{2}{3}k(x_3 - x_2) + c\dot{x}_3 = F_2(t) \end{cases}$$

**FIGURE 4.2**

Mechanical system in Problem 3.

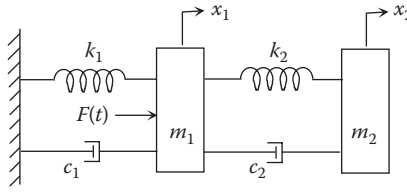


FIGURE 4.3
Mechanical system in Problem 4.

$$\begin{aligned}
 7. \quad & \begin{cases} \frac{1}{2} m \ddot{x}_1 + 2kx_1 - k(x_2 - x_1) - \frac{1}{2} c(\dot{x}_2 - \dot{x}_1) = F_1(t) \\ 2m\ddot{x}_2 - k(x_3 - x_2) + k(x_2 - x_1) + \frac{1}{2} c(\dot{x}_2 - \dot{x}_1) - c(\dot{x}_3 - \dot{x}_2) = F_2(t) \\ m\ddot{x}_3 + 2kx_3 + k(x_3 - x_2) + c(\dot{x}_3 - \dot{x}_2) = 0 \end{cases} \\
 8. \quad & \begin{cases} \ddot{x} = -\frac{1}{4} \dot{x} + F(t) \\ 2\ddot{\phi} = 2\dot{\phi} + \phi - \ddot{x} \end{cases}
 \end{aligned}$$

4.2 State-Space Form

Dynamic systems, ranging from mechanical to fluid and thermal systems, are mathematically modeled as systems of ordinary differential equations (ODEs) of various orders. However, such general form of system model is not very convenient for the purposes of analysis, simulation, and control. One of the most convenient forms is achieved by transforming a system of ODEs of various orders into a larger system of first-order ODEs. The first and most important step toward that goal is the selection of a suitable set of state variables. If there exist n state variables for a specific system, then they are denoted by x_i ($i = 1, 2, \dots, n$).

4.2.1 State Variables, State-Variable Equations, and State Equation

State variables form the smallest set of independent variables that completely describe the state of a system. More exactly, knowledge of the state variables at a fixed reference time t_0 and system inputs at all $t \geq t_0$ translates to knowing the state variables and system outputs at all $t \geq t_0$. State variables are independent; thus, they cannot be expressed as algebraic functions of one another and the system inputs. Moreover, a set of state variables is not unique, so that more than one set may be identified for a dynamic system. Given the mathematical model of a dynamic system, the state variables are determined as follows:

- The number of state variables is the same as the number of initial conditions needed to completely solve the system model.
- The state variables are exactly those variables for which initial conditions are required.

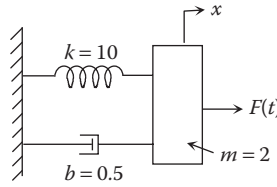


FIGURE 4.4
Mechanical system in Example 4.3.

Example 4.3: State Variables

In the mechanical system shown in Figure 4.4, all parameter values are in consistent physical units. The model of this system is provided by its equation of motion

$$2\ddot{x} + 0.5\dot{x} + 10x = F(t)$$

where the applied force $F(t)$ serves as the system input. Note that we use b and c interchangeably to denote the coefficient of viscous damping.

Two initial conditions, $x(0)$ and $\dot{x}(0)$, are needed to completely solve the differential equation. Therefore, there are two state variables, denoted by x_1 and x_2 . However, the state variables are those variables for which initial conditions are required. Since these variables are x and $\dot{x}$, the state variables are selected as $x_1 = x$ and $x_2 = \dot{x}$.

4.2.1.1 State-Variable Equations

There are as many state-variable equations as there are state variables. Each state-variable equation is a first-order ODE whose left side is the first derivative of a state variable and whose right side is an algebraic function of the state variables, system inputs, and possibly time t . Suppose a dynamic system has n state variables $x_1, x_2, \dots, x_n$ and m inputs $u_1, u_2, \dots, u_m$. Then, the state-variable equations take the generic form

$$\begin{cases} \dot{x}_1 = f_1(x_1, \dots, x_n; u_1, \dots, u_m; t) \\ \dot{x}_2 = f_2(x_1, \dots, x_n; u_1, \dots, u_m; t) \\ \dots \\ \dot{x}_n = f_n(x_1, \dots, x_n; u_1, \dots, u_m; t) \end{cases} \quad (4.3)$$

where $f_1, f_2, \dots, f_n$ are algebraic functions of the state variables and inputs and are generally nonlinear.

Example 4.4: State-Variable Equations (Example 4.3 Continued)

Referring to Example 4.3, since there are two state variables, there must be two state-variable equations in the form

$$\begin{cases} \dot{x}_1 = \dots \\ \dot{x}_2 = \dots \end{cases}$$

Recall that only state variables and inputs may appear on the right side of each ODE. Since $x_1 = x$, we have $\dot{x}_1 = \dot{x}$. However, $\dot{x} = x_2$ and hence $\dot{x}_1 = \dot{x} = x_2$. This means that the

first equation is simply $\dot{x}_1 = x_2$, which is valid because the right side contains only a state variable. Next, we know that $x_2 = \dot{x}$, so that $\dot{x}_2 = \ddot{x}$. However, $\ddot{x}$ is obtained from the equation of motion, as

$$\dot{x}_2 = \ddot{x} \stackrel{\substack{\text{Using the} \\ \text{equation of motion}}}{=} \frac{1}{2}[-0.5\dot{x} - 10x + F(t)] \stackrel{\substack{\text{Use state variables} \\ x_1 = x, x_2 = \dot{x}}}{=} \frac{1}{2}[-0.5x_2 - 10x_1 + F(t)]$$

With this, the state-variable equations can be written as

$$\begin{cases} \dot{x}_1 = x_2 \\ \dot{x}_2 = \frac{1}{2}[-0.5x_2 - 10x_1 + F(t)] \end{cases}$$

Noting that $F(t)$ is the only system input, this agrees with Equation 4.3.

4.2.1.2 State Equation

In general, when at least one of the functions $f_1, f_2, \dots, f_n$ in Equation 4.3 is nonlinear, Equation 4.3 is expressed in vector form, as

$$\dot{\mathbf{x}} = \mathbf{f}(\mathbf{x}, \mathbf{u}, t) \quad (4.4)$$

where

$$\mathbf{x} = \begin{Bmatrix} x_1 \\ x_2 \\ \dots \\ x_n \end{Bmatrix}_{n \times 1}, \quad \mathbf{u} = \begin{Bmatrix} u_1 \\ u_2 \\ \dots \\ u_m \end{Bmatrix}_{m \times 1}, \quad \mathbf{f} = \begin{Bmatrix} f_1 \\ f_2 \\ \dots \\ f_n \end{Bmatrix}_{n \times 1}$$

However, if all elements of a dynamic system are linear, then $f_1, f_2, \dots, f_n$ in Equation 4.3 will be linear combinations of $x_1, x_2, \dots, x_n$ and $u_1, u_2, \dots, u_m$:

$$\begin{cases} \dot{x}_1 = a_{11}x_1 + \dots + a_{1n}x_n + b_{11}u_1 + \dots + b_{1m}u_m \\ \dot{x}_2 = a_{21}x_1 + \dots + a_{2n}x_n + b_{21}u_1 + \dots + b_{2m}u_m \\ \dots \\ \dot{x}_n = a_{n1}x_1 + \dots + a_{nn}x_n + b_{n1}u_1 + \dots + b_{nm}u_m \end{cases} \quad (4.5)$$

Rewriting Equation 4.5 in matrix form yields

$$\begin{Bmatrix} \dot{x}_1 \\ \dot{x}_2 \\ \dots \\ \dot{x}_n \end{Bmatrix} = \begin{bmatrix} a_{11} & a_{12} & \dots & a_{1n} \\ a_{21} & a_{22} & \dots & a_{2n} \\ \dots & \dots & \dots & \dots \\ a_{n1} & a_{n2} & \dots & a_{nn} \end{bmatrix}_{n \times n} \begin{Bmatrix} x_1 \\ x_2 \\ \dots \\ x_n \end{Bmatrix}_{n \times 1} + \begin{bmatrix} b_{11} & b_{12} & \dots & b_{1m} \\ b_{21} & b_{22} & \dots & b_{2m} \\ \dots & \dots & \dots & \dots \\ b_{n1} & b_{n2} & \dots & b_{nm} \end{bmatrix}_{n \times m} \begin{Bmatrix} u_1 \\ u_2 \\ \dots \\ u_m \end{Bmatrix}_{m \times 1}$$

Finally, this can be conveniently expressed as

$$\dot{\mathbf{x}} = \mathbf{Ax} + \mathbf{Bu} \quad (4.6)$$

Equation 4.6 is known as the state equation, where

$$\mathbf{x} = \begin{Bmatrix} x_1 \\ x_2 \\ \dots \\ x_n \end{Bmatrix} = \text{state vector}, \quad \mathbf{u} = \begin{Bmatrix} u_1 \\ u_2 \\ \dots \\ u_m \end{Bmatrix} = \text{input vector}$$

$$\mathbf{A} = \begin{bmatrix} a_{11} & a_{12} & \dots & a_{1n} \\ a_{21} & a_{22} & \dots & a_{2n} \\ \dots & \dots & \dots & \dots \\ a_{n1} & a_{n2} & \dots & a_{nn} \end{bmatrix} = \text{state matrix}, \quad \mathbf{B} = \begin{bmatrix} b_{11} & b_{12} & \dots & b_{1m} \\ b_{21} & b_{22} & \dots & b_{2m} \\ \dots & \dots & \dots & \dots \\ b_{n1} & b_{n2} & \dots & b_{nm} \end{bmatrix} = \text{input matrix}$$

Example 4.5: State Equation (Example 4.4 Continued)

The state-variable equations at the conclusion of Example 4.4 clearly indicate that the system is linear, so they are expressed in matrix form, as

$$\begin{Bmatrix} \dot{x}_1 \\ \dot{x}_2 \end{Bmatrix} = \begin{bmatrix} 0 & 1 \\ -5 & -\frac{1}{4} \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix} + \begin{bmatrix} 0 \\ \frac{1}{2} \end{bmatrix} F(t)$$

Therefore, the state equation is

$$\dot{\mathbf{x}} = \mathbf{A}\mathbf{x} + \mathbf{B}u, \quad \mathbf{x} = \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix}, \quad \mathbf{A} = \begin{bmatrix} 0 & 1 \\ -5 & -\frac{1}{4} \end{bmatrix}, \quad \mathbf{B} = \begin{bmatrix} 0 \\ \frac{1}{2} \end{bmatrix}, \quad u = F(t)$$

Since there is only one input $F(t)$, input vector $\mathbf{u}$ is scalar and denoted by u .

4.2.2 Output Equation and State-Space Form

Consider a dynamic system with state variables $x_1, x_2, \dots, x_n$ and inputs $u_1, u_2, \dots, u_m$, as before. Suppose the system has p outputs $y_1, y_2, \dots, y_p$. Outputs are sometimes called measured outputs, referring to physical quantities that are being measured. Then, the output equations generally appear in the form

$$\begin{cases} y_1 = g_1(x_1, \dots, x_n; u_1, \dots, u_m; t) \\ y_2 = g_2(x_1, \dots, x_n; u_1, \dots, u_m; t) \\ \dots \\ y_p = g_p(x_1, \dots, x_n; u_1, \dots, u_m; t) \end{cases} \quad (4.7)$$

where $g_1, g_2, \dots, g_p$ are algebraic functions of the state variables and inputs and are generally nonlinear.

4.2.2.1 Output Equation

If at least one of the functions $g_1, g_2, \dots, g_p$ in Equation 4.7 is nonlinear, Equation 4.7 is expressed in vector form, as

$$\mathbf{y} = \mathbf{g}(\mathbf{x}, \mathbf{u}, t) \quad (4.8)$$

where

$$\mathbf{y} = \begin{Bmatrix} y_1 \\ y_2 \\ \dots \\ y_p \end{Bmatrix}_{p \times 1}, \quad \mathbf{x} = \begin{Bmatrix} x_1 \\ x_2 \\ \dots \\ x_n \end{Bmatrix}_{n \times 1}, \quad \mathbf{u} = \begin{Bmatrix} u_1 \\ u_2 \\ \dots \\ u_m \end{Bmatrix}_{m \times 1}, \quad \mathbf{g} = \begin{Bmatrix} g_1 \\ g_2 \\ \dots \\ g_p \end{Bmatrix}_{p \times 1}$$

However, if all elements are linear, then $g_1, g_2, \dots, g_p$ in Equation 4.7 are linear combinations of $x_1, x_2, \dots, x_n$ and $u_1, u_2, \dots, u_m$:

$$\begin{cases} y_1 = c_{11}x_1 + \dots + c_{1n}x_n + d_{11}u_1 + \dots + d_{1m}u_m \\ y_2 = c_{21}x_1 + \dots + c_{2n}x_n + d_{21}u_1 + \dots + d_{2m}u_m \\ \dots \\ y_p = c_{p1}x_1 + \dots + c_{pn}x_n + d_{p1}u_1 + \dots + d_{pm}u_m \end{cases}$$

This is conveniently expressed as

$$\mathbf{y} = \mathbf{C}\mathbf{x} + \mathbf{D}\mathbf{u} \quad (4.9)$$

Equation 4.9 is known as the output equation, where $\mathbf{y}$, $\mathbf{x}$, and $\mathbf{u}$ are defined as before, and

$$\mathbf{C} = \begin{bmatrix} c_{11} & c_{12} & \dots & c_{1n} \\ c_{21} & c_{22} & \dots & c_{2n} \\ \dots & \dots & \dots & \dots \\ c_{p1} & c_{p2} & \dots & c_{pn} \end{bmatrix}_{p \times n} = \text{output matrix}$$

$$\mathbf{D} = \begin{bmatrix} d_{11} & d_{12} & \dots & d_{1m} \\ d_{21} & d_{22} & \dots & d_{2m} \\ \dots & \dots & \dots & \dots \\ d_{p1} & d_{p2} & \dots & d_{pm} \end{bmatrix}_{p \times m} = \text{direct transmission matrix}$$

Example 4.6: Output Equation

For the mechanical system studied in Examples 4.3 through 4.5, suppose the output is the velocity $\dot{x}$ of the block. Since there is only one output, the output vector $\mathbf{y}$ is 1×1 and hence denoted by y . The output is $\dot{x}$; therefore, $y = \dot{x}$. However, for this system, we know $\dot{x} = x_2$; thus, $y = x_2$. Finally, Equation 4.9 may be written as

$$y = \begin{bmatrix} 0 & 1 \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix} + 0 \cdot u$$

so that $\mathbf{C} = \begin{bmatrix} 0 & 1 \end{bmatrix}$ and $D = 0$. Note that the direct transmission matrix $\mathbf{D}$ is 1×1 and hence denoted by D .

4.2.2.2 State-Space Form

The combination of the state equation and the output equation is called the state-space form. For a linear system with state variables $x_1, x_2, \dots, x_n$, inputs $u_1, u_2, \dots, u_m$, and outputs $y_1, y_2, \dots, y_p$, the state-space form is

$$\begin{cases} \dot{\mathbf{x}} = \mathbf{A}_{n \times n} \mathbf{x}_{n \times 1} + \mathbf{B}_{n \times m} \mathbf{u}_{m \times 1} \\ \mathbf{y}_{p \times 1} = \mathbf{C}_{p \times n} \mathbf{x}_{n \times 1} + \mathbf{D}_{p \times m} \mathbf{u}_{m \times 1} \end{cases} \quad (4.10)$$

Example 4.7: State-Space Form

In the mechanical system shown in Figure 4.5, all parameter values are in consistent physical units. The equations of motion are derived as

$$\begin{cases} \frac{1}{2} \ddot{x}_1 + 10x_1 - 5(x_2 - x_1) - 0.8(\dot{x}_2 - \dot{x}_1) = F_1 \\ \ddot{x}_2 + 5(x_2 - x_1) + 0.8(\dot{x}_2 - \dot{x}_1) = F_2 \end{cases}$$

Assuming that the (measured) outputs are x_1 and $\dot{x}_1$, obtain the state-space form.

Solution

The system model comprises two second-order ODEs; hence, a total of four initial conditions are needed for complete solution. There are therefore four state variables, selected as

$$x_1 = x_1$$

$$x_2 = x_2$$

$$x_3 = \dot{x}_1$$

$$x_4 = \dot{x}_2$$

When designating state variables, the above-mentioned arrangement is the most commonly used.

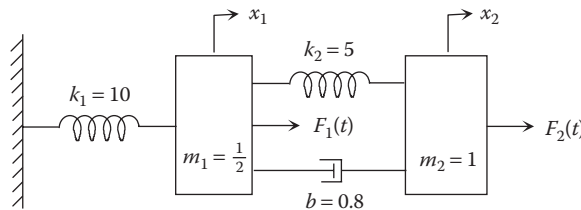


FIGURE 4.5
Mechanical system in Example 4.7.

The state-variable equations are then formed as

$$\begin{cases} \dot{x}_1 = x_3 \\ \dot{x}_2 = x_4 \\ \dot{x}_3 = -30x_1 + 10x_2 + 1.6x_4 - 1.6x_3 + 2F_1 \\ \dot{x}_4 = -5x_2 + 5x_1 - 0.8x_4 + 0.8x_3 + F_2 \end{cases}$$

The state equation is subsequently obtained as

$$\dot{\mathbf{x}} = \mathbf{Ax} + \mathbf{Bu}$$

where

$$\mathbf{x} = \begin{Bmatrix} x_1 \\ x_2 \\ x_3 \\ x_4 \end{Bmatrix}_{4 \times 1}, \quad \mathbf{A} = \begin{bmatrix} 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 1 \\ -30 & 10 & -1.6 & 1.6 \\ 5 & -5 & 0.8 & -0.8 \end{bmatrix}_{4 \times 4}, \quad \mathbf{B} = \begin{bmatrix} 0 & 0 \\ 0 & 0 \\ 2 & 0 \\ 0 & 1 \end{bmatrix}_{4 \times 2}, \quad \mathbf{u} = \begin{Bmatrix} F_1 \\ F_2 \end{Bmatrix}_{2 \times 1}$$

Since the outputs are x_1 and $\dot{x}_1$, we have

$$\mathbf{y} = \begin{Bmatrix} x_1 \\ \dot{x}_1 \end{Bmatrix} = \begin{Bmatrix} x_1 \\ x_3 \end{Bmatrix}$$

As a result, the output equation is

$$\mathbf{y}_{2 \times 1} = \mathbf{C}_{2 \times 4} \mathbf{x}_{4 \times 1} + \mathbf{D}_{2 \times 2} \mathbf{u}_{2 \times 1}$$

where

$$\mathbf{C} = \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 0 & 1 & 0 \end{bmatrix}_{2 \times 4}, \quad \mathbf{D} = \mathbf{0}_{2 \times 2}$$

Finally, combining the state equation and the output equation yields the state-space form.

4.2.2.3 State-Space Form in MATLAB®

The built-in MATLAB function `ss` can be employed to build and store a state-space model. The model can in turn be used for analysis, simulation, and control purposes.

```
>> help ss
```

```
ss Construct state-space model or convert model to state space.
```

Construction:

```
SYS = ss(A,B,C,D) creates an object SYS representing the continuous-time state-space model
```

```
dx/dt = Ax(t) + Bu(t)
```

```
y(t) = Cx(t) + Du(t)
```

Example 4.8: **State-Space Form**

The state-space form in Example 4.7 can be stored in MATLAB under an assigned name, say, `sys`, as follows:

```
% Input matrices A,B,C and D
>> A = [0 0 1 0;0 0 0 1;-30 10 -1.6 1.6;5 -5 0.8 -0.8];
>> B = [0 0;0 0;2 0;0 1]; C = [ 1 0 0 0;0 0 1 0]; D = [0 0;0 0];
>> sys = ss(A,B,C,D)    % state-space form
sys =

a =

      x1      x2      x3      x4
x1      0      0      1      0
x2      0      0      0      1
x3     -30     10     -1.6    1.6
x4      5      -5      0.8   -0.8

b =

      u1      u2
x1      0      0
x2      0      0
x3      2      0
x4      0      1

c =

      x1      x2      x3      x4
y1      1      0      0      0
y2      0      0      1      0

d =

      u1      u2
y1      0      0
y2      0      0
```

Continuous-time state-space model.

4.2.3 Decoupling the State Equation

The state variables are generally coupled through the entries of the state matrix **A**; that is, the state-variable equations cannot be solved independently for the state variables, because the state matrix is not diagonal. However, as we learned in [Chapter 3](#), assuming matrix **A** is not defective, it may be transformed into a diagonal matrix through a coordinate transformation defined by the modal matrix associated with matrix **A**. This process, known as the decoupling of the state equation, is explained as follows.

Consider the state-space form, Equation 4.10. Assume **A** has eigenvalues $\lambda_1, \lambda_2, \dots, \lambda_n$, linearly independent eigenvectors $\mathbf{v}_1, \mathbf{v}_2, \dots, \mathbf{v}_n$, and no generalized eigenvectors. Then, the modal matrix $\mathbf{V} = [\mathbf{v}_1 \ \mathbf{v}_2 \ \dots \ \mathbf{v}_n]_{n \times n}$ diagonalizes matrix **A**, that is,

$$\mathbf{V}^{-1}\mathbf{A}\mathbf{V} = \tilde{\mathbf{D}} = \begin{bmatrix} \lambda_1 & 0 & \dots & 0 \\ 0 & \lambda_2 & 0 & \dots \\ \dots & 0 & \dots & 0 \\ 0 & \dots & 0 & \lambda_n \end{bmatrix}$$

We have changed the notation of **D** to $\tilde{\mathbf{D}}$ to avoid confusion with the direct transmission matrix. Substitute the transformation $\mathbf{x} = \mathbf{V}\tilde{\mathbf{x}}$ into the state-space form to obtain

$$\begin{cases} \dot{\mathbf{V}\tilde{\mathbf{x}}} = \mathbf{A}\mathbf{V}\tilde{\mathbf{x}} + \mathbf{B}\mathbf{u} \\ \mathbf{y} = \mathbf{C}\mathbf{V}\tilde{\mathbf{x}} + \mathbf{D}\mathbf{u} \end{cases} \quad \begin{array}{c} \text{Pre-multiply the state equation} \\ \Rightarrow \\ \text{by } \mathbf{V}^{-1} \end{array} \quad \begin{cases} \mathbf{V}^{-1}\mathbf{V}\dot{\tilde{\mathbf{x}}} = \mathbf{V}^{-1}\mathbf{A}\mathbf{V}\tilde{\mathbf{x}} + \mathbf{V}^{-1}\mathbf{B}\mathbf{u} \\ \mathbf{y} = \mathbf{C}\mathbf{V}\tilde{\mathbf{x}} + \mathbf{D}\mathbf{u} \end{cases}$$

Knowing $\mathbf{V}^{-1}\mathbf{V} = \mathbf{I}$ and $\mathbf{V}^{-1}\mathbf{A}\mathbf{V} = \tilde{\mathbf{D}}$ and denoting $\tilde{\mathbf{B}} = \mathbf{V}^{-1}\mathbf{B}$ and $\tilde{\mathbf{C}} = \mathbf{C}\mathbf{V}$, the above becomes

$$\begin{cases} \dot{\tilde{\mathbf{x}}} = \tilde{\mathbf{D}}\tilde{\mathbf{x}} + \tilde{\mathbf{B}}\mathbf{u} \\ \mathbf{y} = \tilde{\mathbf{C}}\tilde{\mathbf{x}} + \mathbf{D}\mathbf{u} \end{cases} \quad (4.11)$$

Since $\tilde{\mathbf{D}}$ is diagonal, the new state equation in Equation 4.11 is clearly decoupled; each row is a first-order differential equation in one state variable, independent of the others. In the decoupled state-space form, the state vector is $\tilde{\mathbf{x}}$. Equations 4.10 and 4.11 represent the state-space forms of the same system but corresponding to two different state vectors. This is consistent with the fact that the state vector is not unique for a given system model.

Example 4.9: Decoupling the State Equation

The state-space form for a system model is

$$\begin{cases} \dot{\mathbf{x}} = \mathbf{A}\mathbf{x} + \mathbf{B}u \\ y = \mathbf{C}\mathbf{x} + Du \end{cases}$$

where

$$\mathbf{x} = \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix}, \quad \mathbf{A} = \begin{bmatrix} 0 & 1 \\ -\frac{1}{2} & -\frac{3}{2} \end{bmatrix}, \quad \mathbf{B} = \begin{bmatrix} 0 \\ 1 \end{bmatrix}, \quad u = u, \quad \mathbf{C} = \begin{bmatrix} 0 & 1 \end{bmatrix}, \quad D = 0$$

- Find the decoupled state equation, together with the corresponding output equation.
-  Repeat (a) in MATLAB.

Solution

- The eigenvalues and eigenvectors of $\mathbf{A}$ are found as

$$\lambda_{1,2} = -\frac{1}{2}, -1, \quad \mathbf{v}_1 = \begin{Bmatrix} -2 \\ 1 \end{Bmatrix}, \quad \mathbf{v}_2 = \begin{Bmatrix} -1 \\ 1 \end{Bmatrix}$$

Therefore, the modal matrix is formed as $\mathbf{V} = \begin{bmatrix} -2 & -1 \\ 1 & 1 \end{bmatrix}$, and

$$\mathbf{V}^{-1} = \begin{bmatrix} -1 & -1 \\ 1 & 2 \end{bmatrix}, \quad \mathbf{V}^{-1}\mathbf{A}\mathbf{V} = \tilde{\mathbf{D}} = \begin{bmatrix} -\frac{1}{2} & 0 \\ 0 & -1 \end{bmatrix}, \quad \tilde{\mathbf{B}} = \mathbf{V}^{-1}\mathbf{B} = \begin{bmatrix} -1 \\ 2 \end{bmatrix}, \quad \tilde{\mathbf{C}} = \mathbf{C}\mathbf{V} = \begin{bmatrix} 1 & 1 \end{bmatrix}$$

The decoupled state-space form is then obtained as

$$\begin{cases} \dot{\tilde{\mathbf{x}}} = \begin{bmatrix} -\frac{1}{2} & 0 \\ 0 & -1 \end{bmatrix} \tilde{\mathbf{x}} + \begin{bmatrix} -1 \\ 2 \end{bmatrix} u \\ y = \begin{bmatrix} 1 & 1 \end{bmatrix} \tilde{\mathbf{x}} + 0 \cdot u \end{cases}$$

b. 

```

>> A = [0 1;-1/2 -3/2]; B = [0;1]; C = [0 1]; D = 0;
>> [V,Dtilda] = eig(A)
V =
    0.8944    -0.7071
   -0.4472     0.7071
Dtilda =
   -0.5000         0
         0   -1.0000

>> Btilda = V\B; Ctilda = C*V;
>> dec_sys = ss(Dtilda,Btilda,Ctilda,D)    % Define decoupled system
dec_sys =
  a =
      x1      x2
x1   -0.5      0
x2      0     -1

  b =
      u1
x1   2.236
x2   2.828

  c =
      x1      x2
y1  -0.4472     0.7071

  d =
      u1
y1      0

Continuous-time state-space model.

```

Because the "eig" command returns normalized eigenvectors (unit vectors), the ensuing matrix $\tilde{V}$ is different from V obtained in (a), and the subsequent input matrix $\tilde{B}$ and output matrix $\tilde{C}$ do not agree with those in (a). However, the new system is decoupled.

PROBLEM SET 4.2

In Problems 1 through 8, find a suitable set of state variables, derive the state-variable equations, and form the state equation.

1. $m\ddot{x} + b\dot{x} = e^{-2t/3}$, $m, b = \text{const} > 0$
2. $\frac{1}{3}\ddot{x} + \ddot{x} + \frac{1}{2}\dot{x} + 2x = \cos t$
3.
$$\begin{cases} \ddot{x}_1 + \frac{1}{4}\dot{x}_1 + \frac{2}{3}(x_1 - x_2) = F_1(t) \\ 2\dot{x}_2 + x_2 + \frac{2}{3}(x_2 - x_1) = F_2(t) \end{cases}$$
4.
$$\begin{cases} \ddot{x} + \frac{2}{5}\dot{x} + x = 2y \\ 2\dot{y} + \frac{1}{2}y + x = F(t) \end{cases}$$
5.
$$\begin{cases} \ddot{x}_1 + 2\dot{x}_1 + 2(x_1 - x_2) = F(t) \\ \frac{1}{3}x_2 - 2(x_1 - x_2) = 0 \end{cases}$$

6.
$$\begin{cases} \ddot{z}_1 + \dot{z}_1 + \frac{1}{2}k(z_1 - z_2) = e^{-t/3} \\ z_2 - \frac{1}{2}k(z_1 - z_2) = 0 \end{cases}, \quad k = \text{const}$$
7.
$$\begin{cases} 2\ddot{x}_1 + 4\dot{x}_1 - 2\dot{x}_2 + x_1 - x_2 = F_1(t) \\ \ddot{x}_2 - 2\dot{x}_1 + 2\dot{x}_2 - x_1 + x_2 = F_2(t) \end{cases}$$
8.
$$\begin{cases} mL_1^2\ddot{\theta}_1 + (mgL_1 + kL_2^2)\theta_1 - kL_2^2\theta_2 = 0 \\ mL_1^2\ddot{\theta}_2 + (mgL_1 + kL_2^2)\theta_2 - kL_2^2\theta_1 = 0 \end{cases}, \quad m, k, g, L_1, L_2 = \text{const}$$

In Problems 9 and 10, derive the state-variable equations (in vector form) for the given nonlinear system model.

9. $2\ddot{x} + \frac{1}{3}\dot{x}^2 + 2x = e^{-2t/3}$
10.
$$\begin{cases} \ddot{x}_1 + \frac{1}{3}|\dot{x}_1|\dot{x}_1 = x_2^3 \\ \dot{x}_2 = x_1 + 2\sin t \end{cases}$$

Problems 11 through 14 are concerned with the stability of systems. A linear dynamic system is *stable* if the homogeneous solution of its mathematical model, subjected to the prescribed initial conditions, decays. More practically, a linear system is stable if all the eigenvalues of its state matrix have negative real parts; that is, they all lie in the left half plane.

11. Decide whether the system in Problem 1 is stable.
12. Decide whether the system in Problem 5 is stable.
13. Decide whether the system in Problem 4 is stable.
14. Find the range of values of b for which a system described by $\ddot{y} - (b-1)\dot{y} - y = F(t)$ is stable.

In Problems 15 through 18, find the state-space form of the mathematical model.

15.
$$\begin{cases} \ddot{x}_1 + 3\dot{x}_1 + \frac{3}{2}(x_1 - x_2) = 0 \\ 2\dot{x}_2 + x_2 - \frac{3}{2}(x_1 - x_2) = F(t) \end{cases}, \quad \text{outputs are } x_1 \text{ and } \dot{x}_1.$$
16.
$$\begin{cases} \ddot{z}_1 + \frac{3}{4}(z_1 - z_2) + 2\dot{z}_1 + z_1 = F(t) \\ \dot{z}_2 + \frac{3}{4}(z_2 - z_1) = 0 \end{cases}, \quad \text{outputs are } z_1 \text{ and } z_2.$$
17.
$$\begin{cases} 2\ddot{x}_1 + 9(x_1 - x_3) - 0.8(\dot{x}_2 - \dot{x}_1) - 2(x_2 - x_1) = F(t) \\ \ddot{x}_2 + 0.8(\dot{x}_2 - \dot{x}_1) + 2(x_2 - x_1) = 0 \\ 2x_3 = 0.45(x_1 - x_3) \end{cases}, \quad \text{outputs are } x_2 \text{ and } \dot{x}_2.$$
18.
$$\begin{cases} \ddot{x}_1 + \frac{2}{5}(x_1 - x_3) - \frac{3}{5}(\dot{x}_2 - \dot{x}_1) - \frac{1}{2}(x_2 - x_1) = 0 \\ \ddot{x}_2 + \frac{3}{5}(\dot{x}_2 - \dot{x}_1) + \frac{1}{2}(x_2 - x_1) = F(t) \\ \dot{x}_3 = \frac{3}{5}(x_1 - x_3) \end{cases}, \quad \text{outputs are } x_2 \text{ and } x_3.$$
19. A dynamic system is governed by $4\ddot{y} + 8\dot{y} + 3y = f(t)$, where f and y are the system input and output, respectively. Derive the state-space form of the decoupled system.

20. A dynamic system model is derived as

$$\begin{cases} 4\ddot{x}_1 - \dot{x}_1 - 3x_1 + x_2 = f(t) \\ \dot{x}_2 + \frac{1}{3}x_1 + 2x_2 = 0 \end{cases}$$

where the input is f and the outputs are x_2 and $\dot{x}_1$.

- a. Find the state-space form.
- b.  Derive the state-space form of the decoupled system.

4.3 Input–Output Equation and Transfer Function

An I/O equation is a differential equation that relates a system input, a system output, and their time derivatives. If $u(t)$ is the input and $y(t)$ is the output, then the I/O equation is in the general form

$$y^{(n)} + a_1 y^{(n-1)} + \cdots + a_{n-1} \dot{y} + a_n y = b_0 u^{(m)} + b_1 u^{(m-1)} + \cdots + b_{m-1} \dot{u} + b_m u, \quad m \leq n \quad (4.12)$$

where $a_1, \dots, a_n$ and $b_0, b_1, \dots, b_m$ are constants and $y^{(n)} = d^n y / dt^n$. Therefore, a single-input-single-output (SISO) system has only one I/O equation. A multiple-input-multiple-output (MIMO) system, on the other hand, has several I/O equations, one for each pair of I/O. In particular, a system with q inputs and r outputs has a total of qr I/O equations.

4.3.1 Input–Output Equations from the System Model

The generalized coordinates in a system model are normally coupled through the governing equations; for example, the block displacements x_1 and x_2 in the mechanical system of Example 4.7 are coupled through the equations of motion. As a result, deriving one or more I/O equations is usually a difficult task, because it requires that one coordinate be expressed in terms of the other coordinates; for example, in Example 4.7, x_1 would need to be expressed in terms of x_2 , which is not possible due to coupling. In what follows, we present a systematic approach that can be used to obtain I/O equations from the system model. The idea is to take the Laplace transform of the governing equations—assuming zero initial conditions—and eliminate the unwanted variables in the ensuing algebraic system. The new data are subsequently transformed back to time domain and interpreted as one or more differential equations, which in turn are the desired I/O equations.

Example 4.10: Single-Input-Single-Output System

The model for the simple mechanical system in Example 4.3 is derived as

$$2\ddot{x} + 0.5\dot{x} + 10x = F(t)$$

where $F(t)$ is the input and $x(t)$ is the output. Find the I/O equation.

Solution

The equation of motion is already in the form of Equation 4.12; thus, it serves as the I/O equation.

Example 4.11: Multi-Input-Multi-Output System

A system model is described by

$$\begin{cases} \ddot{x}_1 + \dot{x}_1 + x_1 - x_2 = f(t) \\ \ddot{x}_2 + 2\dot{x}_2 - x_1 + x_2 = 0 \end{cases}$$

where f is the input, and x_1 and x_2 are the outputs. Derive all possible I/O equations.

Solution

Since there is one input and two outputs, two I/O equations are expected. Laplace transformation of the model, assuming zero initial conditions, leads to

$$\begin{cases} (s^2 + s + 1)X_1(s) - X_2(s) = F(s) \\ -X_1(s) + (s^2 + 2s + 1)X_2(s) = 0 \end{cases}$$

Since x_1 and x_2 are the outputs, we solve the above system once for $X_1(s)$ and a second time for $X_2(s)$ by using Cramer's rule:

$$X_1(s) = \frac{\begin{vmatrix} F(s) & -1 \\ 0 & s^2 + 2s + 1 \end{vmatrix}}{\begin{vmatrix} s^2 + s + 1 & -1 \\ -1 & s^2 + 2s + 1 \end{vmatrix}} = \frac{(s^2 + 2s + 1)F(s)}{s^4 + 3s^3 + 4s^2 + 3s}$$

$$X_2(s) = \frac{\begin{vmatrix} s^2 + s + 1 & F(s) \\ -1 & 0 \end{vmatrix}}{\begin{vmatrix} s^2 + s + 1 & -1 \\ -1 & s^2 + 2s + 1 \end{vmatrix}} = \frac{F(s)}{s^4 + 3s^3 + 4s^2 + 3s}$$

Cross-multiplication in both cases yields

$$(s^4 + 3s^3 + 4s^2 + 3s)X_1(s) = (s^2 + 2s + 1)F(s)$$

$$(s^4 + 3s^3 + 4s^2 + 3s)X_2(s) = F(s)$$

Interpretation of these two equations in time domain gives the desired possible I/O equations, as

$$x_1^{(4)} + 3\ddot{x}_1 + 4\ddot{x}_1 + 3\dot{x}_1 = \ddot{f} + 2\dot{f} + f, \quad x_2^{(4)} + 3\ddot{x}_2 + 4\ddot{x}_2 + 3\dot{x}_2 = f$$

4.3.2 Transfer Functions from the System Model

A transfer function is defined as the ratio of the Laplace transforms of an output and an input, with the assumption that initial conditions are zero. Therefore, if $y(t)$ is an output and $u(t)$ is an input, the corresponding transfer function is defined as

$$G(s) = \frac{\mathcal{L}\{y(t)\}}{\mathcal{L}\{u(t)\}} = \frac{Y(s)}{U(s)}$$

Consider the general form of I/O equation described by Equation 4.12. Taking the Laplace transform, assuming zero initial conditions, yields

$$(s^n + a_1s^{n-1} + \dots + a_{n-1}s + a_n)Y(s) = (b_0s^m + b_1s^{m-1} + \dots + b_{m-1}s + b_m)U(s)$$

The transfer function is then formed as

$$\frac{Y(s)}{U(s)} = \frac{b_0s^m + b_1s^{m-1} + \dots + b_{m-1}s + b_m}{s^n + a_1s^{n-1} + \dots + a_{n-1}s + a_n}, \quad m \leq n \quad (4.13)$$

As mentioned earlier, there is a transfer function for each I/O pair. A SISO system has therefore only one transfer function, whereas a MIMO system has several, one for each possible I/O pair. If a system has q inputs and r outputs, then there are a total of qr transfer functions, assembled in an $r \times q$ transfer function matrix (also known as a transfer matrix), denoted by $\mathbf{G}(s) = [G_{ij}(s)]$ where $i = 1, 2, \dots, r$ and $j = 1, 2, \dots, q$.

Example 4.12: Transfer Function

A system model is described by

$$\frac{1}{2}\ddot{x} + \dot{x} + 2x = \dot{f} + 2f$$

Assuming that f is the input and x is the output, find the transfer function.

Solution

Assuming zero initial conditions, the Laplace transform of the equation yields

$$(\frac{1}{2}s^2 + s + 2)X(s) = (s + 2)F(s)$$

The transfer function is then formed as

$$\frac{X(s)}{F(s)} = \frac{s + 2}{\frac{1}{2}s^2 + s + 2}$$

Example 4.13: Transfer Matrix

The equations of motion for a mechanical system are given as

$$\begin{cases} \ddot{x}_1 + \dot{x}_1 + 2(x_1 - x_2) = f_1(t) \\ \ddot{x}_2 - 2(x_1 - x_2) = f_2(t) \end{cases}$$

If f_1 and f_2 are the inputs and x_1 and x_2 are the outputs, find the transfer matrix.

Solution

We expect a 2×2 transfer matrix with the following structure:

$$\mathbf{G}(s) = \begin{bmatrix} G_{11}(s) & G_{12}(s) \\ G_{21}(s) & G_{22}(s) \end{bmatrix}_{2 \times 2} = \begin{bmatrix} \left. \frac{X_1(s)}{F_1(s)} \right|_{F_2=0} & \left. \frac{X_1(s)}{F_2(s)} \right|_{F_1=0} \\ \left. \frac{X_2(s)}{F_1(s)} \right|_{F_2=0} & \left. \frac{X_2(s)}{F_2(s)} \right|_{F_1=0} \end{bmatrix}$$

To find the four transfer functions listed earlier, we take the Laplace transform of the governing equations, with zero initial conditions, to obtain

$$\begin{cases} (s^2 + s + 2)X_1(s) - 2X_2(s) = F_1(s) \\ -2X_1(s) + (s^2 + 2)X_2(s) = F_2(s) \end{cases}$$

Solving for $X_1(s)$, we find

$$X_1(s) = \frac{\begin{vmatrix} F_1(s) & -2 \\ F_2(s) & s^2 + 2 \end{vmatrix}}{\begin{vmatrix} s^2 + s + 2 & -2 \\ -2 & s^2 + 2 \end{vmatrix}} = \frac{2}{\Delta(s)} F_2(s) + \frac{s^2 + 2}{\Delta(s)} F_1(s)$$

where

$$\Delta(s) = s^4 + s^3 + 4s^2 + 2s$$

Solving for $X_2(s)$ yields

$$X_2(s) = \frac{\begin{vmatrix} s^2 + s + 2 & F_1(s) \\ -2 & F_2(s) \end{vmatrix}}{\begin{vmatrix} s^2 + s + 2 & -2 \\ -2 & s^2 + 2 \end{vmatrix}} = \frac{2}{\Delta(s)} F_1(s) + \frac{s^2 + s + 2}{\Delta(s)} F_2(s)$$

Note that we do not attempt to cancel terms involving s from the numerator and denominator in each fraction, because valuable information about the system will be lost otherwise. The entries of the 2×2 transfer matrix $\mathbf{G}(s)$ are finally formed as

$$\begin{aligned} G_{11}(s) &= \left. \frac{X_1(s)}{F_1(s)} \right|_{F_2=0} = \frac{s^2 + 2}{\Delta(s)}, & G_{12}(s) &= \left. \frac{X_1(s)}{F_2(s)} \right|_{F_1=0} = \frac{2}{\Delta(s)} \\ G_{21}(s) &= \left. \frac{X_2(s)}{F_1(s)} \right|_{F_2=0} = \frac{2}{\Delta(s)}, & G_{22}(s) &= \left. \frac{X_2(s)}{F_2(s)} \right|_{F_1=0} = \frac{s^2 + s + 2}{\Delta(s)} \end{aligned}$$

Ultimately, the transfer matrix is formed as

$$\mathbf{G}(s) = \begin{bmatrix} \frac{s^2 + 2}{\Delta(s)} & \frac{2}{\Delta(s)} \\ \frac{2}{\Delta(s)} & \frac{s^2 + s + 2}{\Delta(s)} \end{bmatrix}, \quad \Delta(s) = s^4 + s^3 + 4s^2 + 2s$$

PROBLEM SET 4.3

In Problems 1 through 8, find all possible I/O equations.

1. $\begin{cases} \ddot{x}_1 + \dot{x}_1 + \frac{1}{3}(x_1 - x_2) = f(t) \\ \dot{x}_2 - \frac{1}{3}(x_1 - x_2) = 0 \end{cases}, f(t) = \text{input}, x_1, x_2 = \text{outputs}$
2. $\begin{cases} \ddot{x}_1 + 2\dot{x}_1 + \frac{3}{5}(x_1 - x_2) = f_1(t) \\ 3\ddot{x}_2 + \dot{x}_2 - \frac{3}{5}(x_1 - x_2) = f_2(t) \end{cases}, f_1(t), f_2(t) = \text{inputs}, x_2 = \text{output}$

3. $\begin{cases} \ddot{x}_1 + \dot{x}_1 + \frac{1}{3}(x_1 - x_2) = f_1(t) \\ \ddot{x}_2 - \frac{1}{3}(x_1 - x_2) = f_2(t) \end{cases}, f_1(t), f_2(t) = \text{inputs}, x_1 = \text{output}$
 4. $\begin{cases} \ddot{x}_1 + \dot{x}_1 + 2(x_1 - x_2) = 0 \\ \ddot{x}_2 + \dot{x}_2 - 2(x_1 - x_2) = f(t) \end{cases}, f(t) = \text{input}, x_1 = \text{output}$
 5. $\begin{cases} \ddot{\theta}_1 + \frac{2}{3}\dot{\theta}_1 + \frac{1}{3}(\theta_1 - \theta_2) = u(t) \\ \theta_2 - \frac{1}{3}(\theta_1 - \theta_2) = 0 \end{cases}, u(t) = \text{input}, \theta_2 = \text{output}$
 6. $\begin{cases} \ddot{x}_1 + \dot{x}_1 + 3(x_1 - x_2) = f_1(t) \\ \ddot{x}_2 - 3(x_1 - x_2) = f_2(t) \end{cases}, f_1(t), f_2(t) = \text{inputs}, x_1, x_2 = \text{outputs}$
 7. $\begin{cases} \frac{1}{2}\ddot{q}_1 + \dot{q}_1 + 2(q_1 - q_2) + \frac{1}{3}q_1 = v(t) \\ \dot{q}_2 - 2(q_1 - q_2) = 0 \end{cases}, v(t) = \text{input}, q_1, q_2 = \text{outputs}$
 8. $\begin{cases} \ddot{x}_1 + x_1 - x_3 - (\dot{x}_2 - \dot{x}_1) - \frac{1}{2}(x_2 - x_1) = f(t) \\ \ddot{x}_2 + \dot{x}_2 - \dot{x}_1 + \frac{1}{2}(x_2 - x_1) = 0 \\ \dot{x}_3 = x_1 - x_3 \end{cases}, f(t) = \text{input}, x_2, x_3 = \text{outputs}$
 9. A mechanical system model is derived as $m\ddot{x} + b\dot{x} + kx = f(t)$, where $m = 2$, $b = \frac{1}{2}$, $k = 5$, and applied force $f(t) = \frac{2}{5}e^{-t}$, all in consistent physical units. The system is subjected to zero initial conditions. Assuming that x is the output, find the transfer function.
 10. In Problem 9, find the transfer function if $\dot{x}$ is the output.
- In Problems 11 through 14, a system model and its inputs and outputs are provided. Find the appropriate transfer function or matrix.

11. $\begin{cases} \ddot{x}_1 + 0.4\dot{x}_1 + 9(x_1 - x_2) = f(t) \\ 0.6\ddot{x}_2 + 0.8\dot{x}_2 + 9(x_2 - x_1) = 0 \end{cases}, f(t) = \text{input}, x_1 = \text{output}$
12. $\begin{cases} \ddot{q}_1 + \dot{q}_1 + \frac{1}{2}(q_1 - q_2) + q_1 = v(t) \\ \dot{q}_2 - \frac{1}{2}(q_1 - q_2) = 0 \end{cases}, v(t) = \text{input}, q_1 = \text{output}$
13. $\begin{cases} \ddot{x}_1 + x_1 - (x_2 - x_1) - (\dot{x}_2 - \dot{x}_1) = f_1 \\ 2\ddot{x}_2 + (x_2 - x_1) + (\dot{x}_2 - \dot{x}_1) = f_2 \end{cases}, f_1(t), f_2(t) = \text{inputs}, x_1, x_2 = \text{outputs}$
14. $\begin{cases} \ddot{x}_1 + x_1 - x_3 - 2(\dot{x}_2 - \dot{x}_1) = f_1(t) \\ \ddot{x}_2 + 2(\dot{x}_2 - \dot{x}_1) = f_2(t) \\ \dot{x}_3 = x_1 - x_3 \end{cases}, f_1(t), f_2(t) = \text{inputs}, x_1, x_2 = \text{outputs}$
15. The governing equation for an electric circuit is derived as

$$L \frac{di}{dt} + Ri + \frac{1}{C} \int_0^t i(t) dt = v(t)$$

where L , R , and C are the inductance, resistance, and capacitance, respectively, all constants; $i(t)$ is the current; and $v(t)$ is the applied voltage. Initial conditions are assumed to be zero. If $v(t)$ and $i(t)$ are the system input and output, respectively, find the transfer function.

16. Electric charge q and electric current i are related via $i = dq/dt$. In Problem 15, find the transfer function if $v(t)$ and $q(t)$ are the system input and output, respectively.

17. The I/O equation for a dynamic system is given as $3\ddot{x} + 2\dot{x} + 4x = f(t)$, where f and x denote the input and output, respectively.
- Find the system's transfer function.
 - Assuming $f(t)$ is the unit impulse, find the expression for $X(s)$ by using (a).
 - Find the steady-state value x_{ss} via the final-value theorem.
18. The mathematical model of a dynamic system is given as

$$\begin{cases} \ddot{x}_1 + 2\dot{x}_1 + \frac{1}{2}(x_1 - x_2) = f(t) \\ 3\ddot{x}_2 + \dot{x}_2 - \frac{1}{2}(x_1 - x_2) = 0 \end{cases}$$

where f is the input and x_1 and x_2 are the outputs.

- Find the appropriate transfer functions.
 - Assuming $f(t)$ is the unit impulse, find the expressions for $X_1(s)$ and $X_2(s)$ by using (a).
 - Find the steady-state values of x_1 and x_2 by using the final-value theorem.
19. The state-space representation of a system model is described as

$$\begin{cases} \dot{\mathbf{x}} = \mathbf{Ax} + \mathbf{Bu} \\ y = \mathbf{Cx} + \mathbf{Du} \end{cases}$$

where

$$\mathbf{x} = \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix}, \quad \mathbf{A} = \begin{bmatrix} 0 & 1 \\ -2 & -1 \end{bmatrix}, \quad \mathbf{B} = \begin{bmatrix} 0 \\ 1 \end{bmatrix}, \quad \mathbf{C} = \begin{bmatrix} 0 & 1 \end{bmatrix}, \quad \mathbf{D} = 0, \quad u = f$$

Find

- The I/O equation.
 - The transfer function.
20. The state-space representation of a system model is described as

$$\begin{cases} \dot{\mathbf{x}} = \mathbf{Ax} + \mathbf{Bu} \\ \mathbf{y} = \mathbf{Cx} + \mathbf{Du} \end{cases}$$

where

$$\mathbf{x} = \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix}, \quad \mathbf{A} = \begin{bmatrix} 0 & 1 \\ -\frac{1}{2} & -3 \end{bmatrix}, \quad \mathbf{B} = \begin{bmatrix} 0 \\ 1 \end{bmatrix}, \quad \mathbf{C} = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}, \quad \mathbf{D} = \begin{bmatrix} 0 \\ 0 \end{bmatrix}, \quad u = f$$

Find the transfer matrix.

4.4 Relations between State-Space Form, Input–Output Equation, and Transfer Matrix

So far in this chapter, we have learned to derive the state-space form, I/O equation(s), and transfer function(s) directly from the system's mathematical model. In this section, we present two standard procedures to obtain (1) the state-space form from the I/O equation and (2) the transfer matrix from the state-space form.

4.4.1 Input–Output Equation to State-Space Form

Consider the general form of an I/O equation

$$y^{(n)} + a_1 y^{(n-1)} + \dots + a_{n-1} \dot{y} + a_n y = b_0 u^{(n)} + b_1 u^{(n-1)} + \dots + b_{n-1} \dot{u} + b_n u \quad (4.14)$$

where u is the input, y is the output, $a_1, \dots, a_n$ and $b_0, b_1, \dots, b_n$ are constants, and $y^{(n)} = d^n y / dt^n$. Note that this agrees with Equation 4.12, except that the same highest order of differentiation for y and u is now allowed, that is, $m = n$. The goal is to derive the state-space form directly from the I/O equation, Equation 4.14. With the assumption of zero initial conditions, the transfer function is readily obtained as

$$\frac{Y(s)}{U(s)} = \frac{b_0 s^n + b_1 s^{n-1} + \dots + b_{n-1} s + b_n}{s^n + a_1 s^{n-1} + \dots + a_{n-1} s + a_n}$$

Rewrite this expression as

$$\frac{Y(s)}{U(s)} = \frac{Y(s)}{V(s)} \cdot \frac{V(s)}{U(s)} = (b_0 s^n + b_1 s^{n-1} + \dots + b_{n-1} s + b_n) \left(\frac{1}{s^n + a_1 s^{n-1} + \dots + a_{n-1} s + a_n} \right)$$

so that

$$\frac{Y(s)}{V(s)} = b_0 s^n + b_1 s^{n-1} + \dots + b_{n-1} s + b_n, \quad \frac{V(s)}{U(s)} = \frac{1}{s^n + a_1 s^{n-1} + \dots + a_{n-1} s + a_n}$$

Time-domain interpretation of the first one gives

$$y = b_0 v^{(n)} + b_1 v^{(n-1)} + \dots + b_{n-1} \dot{v} + b_n v \quad (4.15)$$

and the second one yields

$$v^{(n)} + a_1 v^{(n-1)} + \dots + a_{n-1} \dot{v} + a_n v = u \quad (4.16)$$

Equation 4.16 is an n th-order differential equation in v ; hence, n initial conditions are needed for complete solution, which implies that there are n state variables, selected as

$$\begin{aligned} x_1 &= v \\ x_2 &= \dot{v} \\ &\dots \\ x_{n-1} &= v^{(n-2)} \\ x_n &= v^{(n-1)} \end{aligned}$$

The resulting state-variable equations are subsequently formed as

$$\begin{aligned}\dot{x}_1 &= x_2 \\ \dot{x}_2 &= x_3 \\ &\vdots \\ \dot{x}_{n-1} &= x_n \\ \dot{x}_n &= -a_n x_1 - a_{n-1} x_2 + \cdots - a_1 x_n + u\end{aligned}$$

Therefore, the state equation is

$$\dot{\mathbf{x}} = \mathbf{Ax} + \mathbf{Bu}$$

where

$$\mathbf{x} = \begin{Bmatrix} x_1 \\ x_2 \\ \vdots \\ x_{n-1} \\ x_n \end{Bmatrix}, \quad \mathbf{A} = \begin{bmatrix} 0 & 1 & 0 & \cdots & 0 \\ 0 & 0 & 1 & \cdots & 0 \\ \vdots & \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & 0 & \cdots & 1 \\ -a_n & -a_{n-1} & -a_{n-2} & \cdots & -a_1 \end{bmatrix}_{n \times n}, \quad \mathbf{B} = \begin{bmatrix} 0 \\ 0 \\ \vdots \\ 0 \\ 1 \end{bmatrix}_{n \times 1}, \quad u = u \quad (4.17)$$

Lower companion matrix

The state matrix in Equation 4.17 is called the lower companion matrix. The output is given by Equation 4.15. Using the state variables in Equation 4.15, we find

$$\begin{aligned}y &= b_0 \ddot{v}^{(n)} + b_1 \ddot{v}^{(n-1)} + \cdots + b_{n-1} \dot{v} + b_n v \\ &= b_0 \dot{x}_n + b_1 x_n + \cdots + b_{n-1} x_2 + b_n x_1\end{aligned}$$

Since the output equation cannot contain $\dot{x}_n$, we substitute for $\dot{x}_n$ by using the last relation in the state-variable equations. The result is

$$\begin{aligned}y &= b_0(-a_n x_1 - a_{n-1} x_2 + \cdots - a_1 x_n + u) + b_1 x_n + \cdots + b_{n-1} x_2 + b_n x_1 \\ &\stackrel{\text{collect like terms}}{=} (-b_0 a_n + b_n) x_1 + (-b_0 a_{n-1} + b_{n-1}) x_2 + \cdots + (-b_0 a_1 + b_1) x_n + b_0 u\end{aligned}$$

Finally, the output equation is obtained as

$$y = \mathbf{Cx} + Du$$

where

$$\mathbf{C} = \begin{bmatrix} -b_0 a_n + b_n & -b_0 a_{n-1} + b_{n-1} & \cdots & -b_0 a_1 + b_1 \end{bmatrix}_{1 \times n}, \quad D = b_0 \quad (4.18)$$

Equations 4.17 and 4.18 describe all four matrices involved in the state-space form.

Example 4.14: Input/Output Equation to State-Space Form

Find the state-space form corresponding to the I/O equation

$$3\ddot{y} + \ddot{y} + \dot{y} + 2y = \ddot{u} + \frac{1}{2}\dot{u} + u$$

Solution

Divide the I/O equation by 3 to obtain

$$\ddot{y} + \frac{1}{3}\ddot{y} + \frac{1}{3}\dot{y} + \frac{2}{3}y = \frac{1}{3}\ddot{u} + \frac{1}{6}\dot{u} + \frac{1}{3}u$$

Comparing with Equation 4.14, we have $n = 3$, $a_1 = \frac{1}{3}$, $a_2 = \frac{1}{3}$, $a_3 = \frac{2}{3}$, $b_0 = 0$, $b_1 = \frac{1}{3}$, $b_2 = \frac{1}{6}$, and $b_3 = \frac{1}{3}$. By Equation 4.17,

$$\mathbf{x} = \begin{Bmatrix} x_1 \\ x_2 \\ x_3 \end{Bmatrix}, \quad \mathbf{A} = \begin{bmatrix} 0 & 1 & 0 \\ 0 & 0 & 1 \\ -\frac{2}{3} & -\frac{1}{3} & -\frac{1}{3} \end{bmatrix}, \quad \mathbf{B} = \begin{bmatrix} 0 \\ 0 \\ 1 \end{bmatrix}, \quad u = u$$

By Equation 4.18,

$$\mathbf{C} = \begin{bmatrix} \frac{1}{3} & \frac{1}{6} & \frac{1}{3} \end{bmatrix}, \quad D = 0$$

Therefore, the state-space form is obtained as

$$\begin{cases} \dot{\mathbf{x}} = \begin{bmatrix} 0 & 1 & 0 \\ 0 & 0 & 1 \\ -\frac{2}{3} & -\frac{1}{3} & -\frac{1}{3} \end{bmatrix} \mathbf{x} + \begin{bmatrix} 0 \\ 0 \\ 1 \end{bmatrix} u \\ y = \begin{bmatrix} \frac{1}{3} & \frac{1}{6} & \frac{1}{3} \end{bmatrix} \mathbf{x} \end{cases}$$

4.4.1.1 Controller Canonical Form

The process of obtaining the state-space form from the I/O equation can be handled in MATLAB by using the built-in "tf2ss" command.

```
>> help tf2ss
tf2ss Transfer function to state-space conversion.
[A,B,C,D] = tf2ss(NUM,DEN) calculates the state-space
representation:
    x = Ax + Bu
    y = Cx + Du
of the system:
          NUM(s)
H(s) =  -----
          DEN(s)
from a single input. Vector DEN must contain the coefficients of
the denominator in descending powers of s. Matrix NUM must
contain the numerator coefficients with as many rows as there are
outputs y. The A,B,C,D matrices are returned in controller
canonical form. This calculation also works for discrete systems.
```

The `tf2ss` command calls for the transfer function, which is available from the I/O equation, and returns the state-space form in *controller canonical form*. This form is different from that in Equations 4.17 and 4.18, because it is based on the state variables being selected in the reverse order of what we have become accustomed to. It can be shown (see Problem Set 4.4) that the selection of state variables as

$$\begin{aligned}x_1 &= v^{(n-1)} \\x_2 &= v^{(n-2)} \\&\dots \\x_{n-1} &= \dot{v} \\x_n &= v\end{aligned}$$

leads to the state-space representation in controller canonical form:

$$\begin{cases} \dot{\mathbf{x}} = \mathbf{A}\mathbf{x} + \mathbf{B}u \\ \mathbf{y} = \mathbf{C}\mathbf{x} + \mathbf{D}u \end{cases}$$

where

$$\mathbf{x} = \begin{Bmatrix} x_1 \\ x_2 \\ \dots \\ x_{n-1} \\ x_n \end{Bmatrix}, \quad \mathbf{A} = \begin{bmatrix} -a_1 & -a_2 & \dots & -a_{n-2} & -a_{n-1} & -a_n \\ 1 & 0 & \dots & 0 & 0 & 0 \\ \dots & \dots & \dots & \dots & \dots & \dots \\ 0 & 0 & \dots & 1 & 0 & 0 \\ 0 & 0 & \dots & 0 & 1 & 0 \end{bmatrix}_{n \times n}, \quad \mathbf{B} = \begin{bmatrix} 1 \\ 0 \\ \dots \\ 0 \\ 0 \end{bmatrix}_{n \times 1}, \quad u = u \quad (4.19)$$

Upper companion matrix

and

$$\mathbf{C} = [-b_0a_1 + b_1 \quad -b_0a_2 + b_2 \quad \dots \quad -b_0a_n + b_n]_{1 \times n}, \quad D = b_0 \quad (4.20)$$

The state matrix in the controller canonical form is known as the upper companion matrix.

Example 4.15: Input/Output Equation to Controller Canonical State-Space Form

A system's I/O equation is provided as

$$\ddot{y} + 2\dot{y} + \frac{1}{2}y = \ddot{u} + \ddot{u} + 3\dot{u} + 2u$$

- Find the state-space representation in controller canonical form.
- Confirm the results of (a) in MATLAB.

Solution

- Comparing the I/O equation with Equation 4.14, we find

$$n = 3, a_1 = 2, a_2 = 1, a_3 = \frac{1}{2}, b_0 = 1, b_1 = 1, b_2 = 3, \text{ and } b_3 = 2$$

Selecting the state variables as $x_1 = \ddot{y}$, $x_2 = \dot{y}$, and $x_3 = y$, Equations 4.19 and 4.20 yield

$$\begin{cases} \dot{\mathbf{x}} = \begin{bmatrix} -2 & -1 & -\frac{1}{2} \\ 1 & 0 & 0 \\ 0 & 1 & 0 \end{bmatrix} \mathbf{x} + \begin{bmatrix} 1 \\ 0 \\ 0 \end{bmatrix} u \\ y = \begin{bmatrix} -1 & 2 & \frac{3}{2} \end{bmatrix} \mathbf{x} + 1 \cdot u \end{cases}$$

b.  The transfer function is directly obtained from the I/O equation, as

$$\frac{Y(s)}{U(s)} = \frac{s^3 + s^2 + 3s + 2}{s^3 + 2s^2 + s + \frac{1}{2}}$$

```
>> Num = [1 1 3 2]; % Define numerator
>> Den = [1 2 1 1/2]; % Define denominator
>> [A,B,C,D] = tf2ss(Num,Den)
A =
    -2.0000    -1.0000    -0.5000
     1.0000         0         0
         0         1.0000         0
B =
     1
     0
     0
C =
    -1.0000     2.0000     1.5000
D =
     1
```

4.4.2 State-Space Form to Transfer Matrix

The transfer function (for SISO systems) or transfer matrix (for MIMO systems) can be systematically derived from the state-space form. Consider the state-space form, as in Equation 4.10

$$\begin{cases} \dot{\mathbf{x}} = \mathbf{A}_{n \times n} \mathbf{x}_{n \times 1} + \mathbf{B}_{n \times m} \mathbf{u}_{m \times 1} \\ \mathbf{y}_{p \times 1} = \mathbf{C}_{p \times n} \mathbf{x}_{n \times 1} + \mathbf{D}_{p \times m} \mathbf{u}_{m \times 1} \end{cases}$$

Since the system has m inputs and p outputs, there are a total of mp transfer functions, and the transfer matrix $\mathbf{G}(s)$ is $p \times m$ and derived as follows. First, we note that the Laplace transform of a vector such as $\mathbf{x}$ is handled as

$$\mathbf{x} = \begin{Bmatrix} x_1 \\ x_2 \\ \vdots \\ x_n \end{Bmatrix} \xrightarrow{\text{Laplace transform}} \mathcal{L}\{\mathbf{x}\} = \mathbf{X}(s) = \begin{Bmatrix} \mathcal{L}\{x_1\} \\ \mathcal{L}\{x_2\} \\ \vdots \\ \mathcal{L}\{x_n\} \end{Bmatrix}$$

Assuming zero initial state vector, $\mathbf{x}(0) = \mathbf{0}_{n \times 1}$, Laplace transformation of the state-space form yields

$$\begin{cases} s\mathbf{X}(s) = \mathbf{A}\mathbf{X}(s) + \mathbf{B}\mathbf{U}(s) \\ \mathbf{Y}(s) = \mathbf{C}\mathbf{X}(s) + \mathbf{D}\mathbf{U}(s) \end{cases}$$

The first equation is manipulated as

$$(s\mathbf{I} - \mathbf{A})\mathbf{X}(s) = \mathbf{B}\mathbf{U}(s) \quad \begin{array}{c} \text{Pre-multiply by} \\ \Rightarrow \\ (s\mathbf{I} - \mathbf{A})^{-1} \end{array} \quad \mathbf{X}(s) = (s\mathbf{I} - \mathbf{A})^{-1}\mathbf{B}\mathbf{U}(s)$$

Inserting this into the second equation results in

$$\begin{aligned} \mathbf{Y}(s) &= \mathbf{C}(s\mathbf{I} - \mathbf{A})^{-1}\mathbf{B}\mathbf{U}(s) + \mathbf{D}\mathbf{U}(s) \\ &= [\mathbf{C}(s\mathbf{I} - \mathbf{A})^{-1}\mathbf{B} + \mathbf{D}]\mathbf{U}(s) \end{aligned}$$

For a SISO system with input u and output y , the transfer function is $G(s) = Y(s)/U(s)$, so that $Y(s) = G(s)U(s)$. Extending this idea to MIMO systems, with input vector $\mathbf{u}$ and output vector $\mathbf{y}$, we have $\mathbf{Y}(s) = \mathbf{G}(s)\mathbf{U}(s)$. However, earlier analysis led to $\mathbf{Y}(s) = [\mathbf{C}(s\mathbf{I} - \mathbf{A})^{-1}\mathbf{B} + \mathbf{D}]\mathbf{U}(s)$. Comparison of these last two relations suggests that the transfer matrix is defined as

$$\mathbf{G}(s) = \mathbf{C}(s\mathbf{I} - \mathbf{A})^{-1}\mathbf{B} + \mathbf{D} \quad (4.21)$$

4.4.2.1 MATLAB Command "ss2tf"

The process of obtaining the transfer matrix from the state-space form is handled in MATLAB by using the built-in "ss2tf" command.

```
>> help ss2tf
ss2tf State-space to transfer function conversion.
      [NUM,DEN] = ss2tf(A,B,C,D,iu) calculates the transfer function:
```

$$H(s) = \frac{\text{NUM}(s)}{\text{DEN}(s)} = \frac{\text{---}^{-1}}{\text{---}} = \mathbf{C}(s\mathbf{I} - \mathbf{A})^{-1}\mathbf{B} + \mathbf{D}$$

of the system:

$$\begin{aligned} \dot{\mathbf{x}} &= \mathbf{A}\mathbf{x} + \mathbf{B}\mathbf{u} \\ \mathbf{y} &= \mathbf{C}\mathbf{x} + \mathbf{D}\mathbf{u} \end{aligned}$$

from the i u'th input. Vector DEN contains the coefficients of the denominator in descending powers of s . The numerator coefficients are returned in matrix NUM with as many rows as there are outputs y .

Example 4.16: Single-Input-Single-Output System

A system's state-space representation is

$$\begin{cases} \dot{\mathbf{x}} = \mathbf{A}\mathbf{x} + \mathbf{B}u \\ y = \mathbf{C}\mathbf{x} + \mathbf{D}u \end{cases}$$

where

$$\mathbf{x} = \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix}, \quad \mathbf{A} = \begin{bmatrix} 0 & 1 \\ -1 & -\frac{2}{3} \end{bmatrix}, \quad \mathbf{B} = \begin{bmatrix} 0 \\ 1 \end{bmatrix}, \quad \mathbf{C} = \begin{bmatrix} 1 & \frac{1}{2} \end{bmatrix}, \quad \mathbf{D} = 0, \quad u = u$$

- Find the transfer function.
-  Confirm (a) by using `ss2tf`.

Solution

- Since both u and y are 1×1 , the system is SISO; hence, there is only one transfer function. Given the matrix sizes, $\mathbf{G}(s)$ in Equation 4.21 is easily confirmed to be 1×1 and simply denoted by $G(s)$. Noting $\mathbf{D} = 0$, Equation 4.21 reduces to

$$G(s) = \mathbf{C}(s\mathbf{I} - \mathbf{A})^{-1}\mathbf{B}$$

Using the adjoint matrix ([Chapter 3](#)), we find

$$(s\mathbf{I} - \mathbf{A})^{-1} = \frac{1}{s^2 + \frac{2}{3}s + 1} \begin{bmatrix} s + \frac{2}{3} & 1 \\ -1 & s \end{bmatrix}$$

With this, the transfer function is obtained as

$$G(s) = \mathbf{C}(s\mathbf{I} - \mathbf{A})^{-1}\mathbf{B} = \begin{bmatrix} 1 & \frac{1}{2} \end{bmatrix} \frac{1}{s^2 + \frac{2}{3}s + 1} \begin{bmatrix} s + \frac{2}{3} & 1 \\ -1 & s \end{bmatrix} \begin{bmatrix} 0 \\ 1 \end{bmatrix} = \frac{\frac{1}{2}s + 1}{s^2 + \frac{2}{3}s + 1}$$

- 

```
>> A = [0 1; -1 -2/3]; B = [0; 1]; C = [1 1/2]; D = 0;
>> [Num, Den] = ss2tf(A,B,C,D)
Num =
      0      0.5000      1.0000      % Num = (1/2)s + 1
Den =
      1.0000      0.6667      1.0000      % Den = s^2 + (2/3)s + 1
```

Example 4.17: Multi-Input-Multi-Output System

The state-space form for a system model is

$$\begin{cases} \dot{\mathbf{x}} = \mathbf{A}\mathbf{x} + \mathbf{B}\mathbf{u} \\ \mathbf{y} = \mathbf{C}\mathbf{x} + \mathbf{D}\mathbf{u} \end{cases}$$

where

$$\mathbf{x} = \begin{Bmatrix} x_1 \\ x_2 \\ x_3 \end{Bmatrix}, \quad \mathbf{A} = \begin{bmatrix} 0 & 1 & 0 \\ 0 & 1 & -1 \\ -\frac{1}{3} & -2 & -\frac{2}{3} \end{bmatrix}, \quad \mathbf{B} = \begin{bmatrix} 0 & 0 \\ 1 & 0 \\ 0 & \frac{1}{2} \end{bmatrix}, \quad \mathbf{C} = \begin{bmatrix} 0 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix}, \quad \mathbf{D} = \mathbf{0}_{2 \times 2}$$

Find the transfer matrix by using `ss2tf`.

Solution

It is observed that both $\mathbf{y}$ and $\mathbf{u}$ are 2×1 ; hence, there are two outputs and two inputs and the transfer matrix $\mathbf{G}(s)$ is 2×2 .

```
>> A = [0 1 0; 0 1 -1; -1/3 -2 -2/3]; B = [0 0; 1 0; 0 1/2];
>> C = [0 1 0; 0 0 1]; D = zeros(2,2);
>> [Num1, Den1] = ss2tf(A,B,C,D,1) % Contribution by input 1
Num1 =
    0    1.0000    0.6667    0    % s^2 + (2/3)s
    0         0   -2.0000   -0.3333 % -2s - 1/3
Den1 =
    1.0000   -0.3333   -2.6667   -0.3333 % s^3 - (1/3)s^2 - (8/3)s - 1/3
>> [Num2, Den2] = ss2tf(A,B,C,D,2) % Contribution by input 2
Num2 =
    0         0   -0.5000    0    % -(1/2)s
    0    0.5000   -0.5000    0    % (1/2)s^2 - (1/2)s
Den2 =
    1.0000   -0.3333   -2.6667   -0.3333 % s^3 - (1/3)s^2 - (8/3)s - 1/3
```

The two rows returned in Num1 translate to $s^2 + \frac{2}{3}s$ and $-2s - \frac{1}{3}$. These represent the numerators in the two transfer functions relating the first input to the two outputs. Recall from previous work that these two transfer functions occupy the first column of the transfer matrix. The two rows returned in Num2 translate to $-\frac{1}{2}s$ and $\frac{1}{2}s^2 - \frac{1}{2}s$, which are the numerators in the two transfer functions relating the second input to the two outputs. These two transfer functions occupy the second column of the transfer matrix. Therefore, the transfer matrix is formed as

$$\mathbf{G}(s) = \frac{1}{s^3 - \frac{1}{3}s^2 - \frac{8}{3}s - \frac{1}{3}} \begin{bmatrix} s^2 + \frac{2}{3}s & -\frac{1}{2}s \\ -2s - \frac{1}{3} & \frac{1}{2}s^2 - \frac{1}{2}s \end{bmatrix}$$

PROBLEM SET 4.4

In Problem 1 through 6, find the state-space form directly from the I/O equation.

1. $\frac{1}{2}\ddot{y} + \dot{y} + y = \dot{u}$
2. $3\ddot{y} + \dot{y} + 2y = 3\ddot{u} + 2u$
3. $\ddot{y} + \frac{2}{5}\dot{y} + \dot{y} + 2y = \ddot{u} + 2\dot{u}$
4. $\frac{2}{3}\ddot{y} + \dot{y} + \frac{3}{2}y = \frac{1}{3}\ddot{u} + \dot{u} + u$
5. $2y^{(4)} + \ddot{y} + 2\dot{y} + y = \ddot{u} + 2\dot{u} + u$
6. $y^{(4)} + \ddot{y} + \frac{1}{4}\dot{y} + y = 4u$
7. Derive the state-space representation in controller canonical form, as described by Equations 4.19 and 4.20.
8. A system's I/O equation is provided as

$$\frac{5}{6}y^{(4)} + \ddot{y} + \frac{1}{6}\dot{y} + \frac{1}{2}y = \ddot{u} + \dot{u} + \frac{1}{3}u$$

- a. Find the state-space representation in controller canonical form.
- b.  Confirm the results of (a) in MATLAB.

In Problems 9 through 12, given the transfer function $Y(s)/U(s)$, find

- The I/O equation.
- The state-space form directly from the I/O equation in (a).

9. $\frac{s + \frac{3}{4}}{s^3 + \frac{1}{3}s + 1}$

10. $\frac{\frac{1}{5}s^2 + 1}{s^2 + \frac{3}{5}s + \frac{1}{10}}$

11. $\frac{s^3 + 2s + 1}{\frac{2}{3}s^3 + s^2 + s + 1}$

12. $\frac{s^3 + 2s}{2s^4 + s^2 + 3s + 2}$

13. The state-variable equations and the output equation for a dynamic system are given as

$$\begin{cases} \dot{x}_1 = x_2 \\ \dot{x}_2 = -x_2 - x_1 + u \end{cases}, \quad y = \frac{1}{2}x_1 + x_2$$

Find the transfer function (or matrix) by determining the Laplace transforms of x_1 and x_2 in the state-variable equations and using them in the Laplace transform of the output equation.

14. Repeat Problem 13 for

$$\begin{cases} \dot{x}_1 = x_2 \\ \dot{x}_2 = -\frac{1}{2}x_2 - x_1 + \frac{1}{2}u \end{cases}, \quad \mathbf{y} = \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix}$$

In Problems 15 through 19, the matrices in the state-space form of a system model are given.

- Find the transfer function (or transfer matrix) by using Equation 4.21.
- Verify the result of (a) by using the `ss2tf` command.

15. $\mathbf{A} = \begin{bmatrix} 0 & 1 \\ -2 & -\frac{1}{7} \end{bmatrix}$, $\mathbf{B} = \begin{bmatrix} 0 \\ 2 \end{bmatrix}$, $\mathbf{C} = \begin{bmatrix} \frac{1}{2} & 1 \end{bmatrix}$, $D = 0$

16. $\mathbf{A} = \begin{bmatrix} 0 & 1 \\ -1 & -2 \end{bmatrix}$, $\mathbf{B} = \begin{bmatrix} 0 \\ \frac{2}{3} \end{bmatrix}$, $\mathbf{C} = \begin{bmatrix} 0 & 1 \end{bmatrix}$, $D = \frac{1}{3}$

17. $\mathbf{A} = \begin{bmatrix} 0 & 1 & 0 \\ 0 & 0 & 1 \\ -\frac{2}{3} & -\frac{1}{3} & -1 \end{bmatrix}$, $\mathbf{B} = \begin{bmatrix} 0 \\ 0 \\ \frac{1}{2} \end{bmatrix}$, $\mathbf{C} = \begin{bmatrix} 1 & 0 & 0 \\ 0 & 0 & 1 \end{bmatrix}$, $\mathbf{D} = \mathbf{0}_{2 \times 1}$

$$18. \mathbf{A} = \begin{bmatrix} 0 & 1 & 0 \\ 0 & 0 & 1 \\ -2 & -1 & -2 \end{bmatrix}, \mathbf{B} = \begin{bmatrix} 0 & 0 \\ 1 & 0 \\ 0 & \frac{1}{4} \end{bmatrix}, \mathbf{C} = \begin{bmatrix} 0 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix}, \mathbf{D} = \begin{bmatrix} \frac{1}{2} & 0 \\ 0 & 0 \end{bmatrix}$$

$$19. \mathbf{A} = \begin{bmatrix} 0 & 1 & 0 \\ 0 & 0 & 1 \\ -1 & 0 & -1 \end{bmatrix}, \mathbf{B} = \begin{bmatrix} 0 & 0 \\ 1 & 0 \\ 0 & 2 \end{bmatrix}, \mathbf{C} = \begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \end{bmatrix}, \mathbf{D} = \begin{bmatrix} 0 & 0 \\ 0 & 1 \end{bmatrix}$$

20. A dynamic system model has state variables x_1 and x_2 and input u . The two I/O equations are derived as

$$\ddot{x}_1 + \dot{x}_1 + 2x_1 = \dot{u} + \frac{1}{3}u, \quad \ddot{x}_2 + \dot{x}_2 + 2x_2 = 2u$$

Find the state-space representation for this model and subsequently the transfer matrix.

4.5 Block Diagram Representation

A block diagram that represents a dynamic system is an interconnection of blocks, each block corresponding to an operation carried out by a component in such a way that the block diagram as a whole agrees with the mathematical model of the system. Each block is identified with a transfer function $G(s) = O(s)/I(s)$, also called the gain of the block, as shown in [Figure 4.6](#). Therefore, the output of the block is

$$O(s) = G(s)I(s)$$

4.5.1 Block Diagram Operations

The principal operations in block diagrams include signal amplification, algebraic summation of signals, integration of signals, replacing series and parallel block combinations with equivalent single blocks, and treatment of loops.

4.5.1.1 Summing Junction

The output of a summing junction (or summer) is the algebraic sum of signals entering the summing junction. Each signal is accompanied by a positive or negative sign; see [Figure 4.7](#). A summing junction may have as many inputs (with the same units) as desired but only one single output.

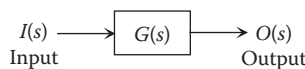


FIGURE 4.6
Schematic of a transfer function block.

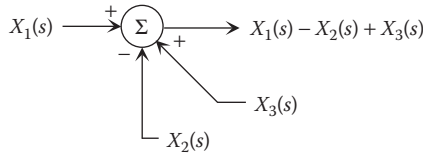


FIGURE 4.7
Summing junction.

4.5.1.2 Series Combinations of Blocks

Consider two blocks with transfer functions $G_1(s)$ and $G_2(s)$ in a series combination, as in [Figure 4.8](#). The block $G_1(s)$ has input $U(s)$ and output $X(s)$, which is the input to the block $G_2(s)$. The output of block $G_2(s)$ is $Y(s)$. Therefore, the overall input is $U(s)$, whereas the overall output is $Y(s)$. The goal is to replace this arrangement with a single block that has $U(s)$ as input and $Y(s)$ as output, that is, a single block with transfer function $Y(s)/U(s)$.

The output of the second block is $Y(s) = G_2(s)X(s)$. However, since $X(s)$ is the output of the first block and is given by $X(s) = G_1(s)U(s)$, we have

$$Y(s) = G_2(s)X(s) = G_2(s)[G_1(s)U(s)] = [G_1(s)G_2(s)]U(s)$$

Therefore,

$$\frac{Y(s)}{U(s)} = G_1(s)G_2(s)$$

As a result, the series configuration in [Figure 4.8](#) can be replaced with a single block $G_1(s)G_2(s)$ with overall input $U(s)$ and overall output $Y(s)$, as shown in [Figure 4.9](#).

Example 4.18: Blocks in Series

Consider two blocks in a series connection with transfer functions

$$G_1(s) = \frac{s}{s^2 + 2}, \quad G_2(s) = \frac{s + \frac{1}{2}}{s^2 + 2s}$$

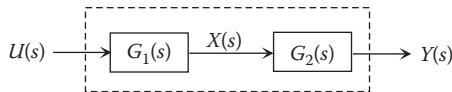


FIGURE 4.8
Two blocks in series.

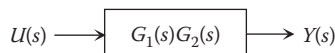


FIGURE 4.9
A single block replacing two blocks in series.

The transfer function of the equivalent single block is determined as

$$G_1(s)G_2(s) = \frac{s(s + \frac{1}{2})}{(s^2 + 2)(s^2 + 2s)}$$

This can be handled in MATLAB as follows:

```
>> Num1 = [1 0]; Den1 = [1 0 2]; Num2 = [1 1/2]; Den2 = [1 2 0];
>> sysG1 = tf(Num1, Den1);      % Define system w/ transfer function G1(s)
>> sysG2 = tf(Num2, Den2);      % Define system w/ transfer function G2(s)
>> sysEq = series(sysG1,sysG2)  % Find the equivalent single TF
```

```
sysEq =
```

$$\frac{s^2 + 0.5 s}{s^4 + 2 s^3 + 2 s^2 + 4 s}$$

```
Continuous-time transfer function.
```

4.5.1.3 Parallel Combinations of Blocks

Consider two blocks with transfer functions $G_1(s)$ and $G_2(s)$ in a parallel combination, as shown in Figure 4.10. Once again, the objective is to replace the arrangement with a single block that has $U(s)$ as input and $Y(s)$ as output, that is, a single block with transfer function $Y(s)/U(s)$. Note that point B in Figure 4.10 is called a branch point. The outputs of the two blocks, labeled $Y_1(s)$ and $Y_2(s)$, are simply $Y_1(s) = G_1(s)U(s)$ and $Y_2(s) = G_2(s)U(s)$, respectively. These two are the inputs to the summing junction, whose output is then calculated as

$$Y(s) = Y_1(s) + Y_2(s) = G_1(s)U(s) + G_2(s)U(s) = [G_1(s) + G_2(s)]U(s)$$

Therefore,

$$\frac{Y(s)}{U(s)} = G_1(s) + G_2(s)$$

The parallel configuration in Figure 4.10 can thus be replaced with a single block $G_1(s) + G_2(s)$, overall input $U(s)$, and overall output $Y(s)$, as shown in Figure 4.11.

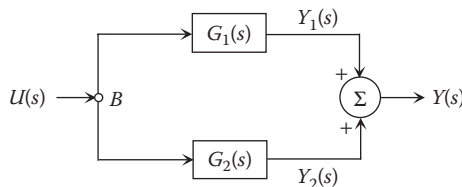
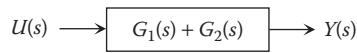


FIGURE 4.10
Two blocks in parallel.

**FIGURE 4.11**

A single block replacing two blocks in parallel.

Example 4.19: Blocks in Parallel

Consider two blocks in a parallel connection with transfer functions

$$G_1(s) = \frac{1}{2s+1}, \quad G_2(s) = \frac{s + \frac{1}{3}}{s^2 + 2s}$$

The transfer function of the equivalent single block is determined as

$$G_1(s) + G_2(s) = \frac{1}{2s+1} + \frac{s + \frac{1}{3}}{s^2 + 2s} = \frac{3s^2 + \frac{11}{3}s + \frac{1}{3}}{(2s+1)(s^2 + 2s)}$$

This can be handled in MATLAB as follows:

```
>> sysG1 = tf([1], [2 1]);
>> sysG2 = tf([1 1/3], [1 2 0]);
>> sysEq = parallel(sysG1, sysG2)
```

```
sysEq =
```

$$\frac{3s^2 + 3.667s + 0.3333}{2s^3 + 5s^2 + 2s}$$

```
Continuous-time transfer function.
```

4.5.1.4 Integrator

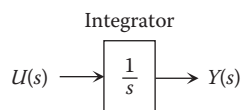
An integrator integrates a signal $u(t)$ from initial time 0 to the current time t to produce

$$y(t) = \int_0^t u(t) dt$$

Laplace transformation of this relation yields

$$Y(s) = \frac{1}{s} U(s) \quad \begin{array}{c} \text{Integrator} \\ \Rightarrow \\ \text{transfer function} \end{array} \quad \frac{Y(s)}{U(s)} = \frac{1}{s}$$

Therefore, an integrator is represented by a single block with transfer function $1/s$, as shown in [Figure 4.12](#).

**FIGURE 4.12**

Integrator block.

4.5.1.5 Closed-Loop Systems

A closed-loop (or feedback) system is depicted in [Figure 4.13](#). The input is $U(s)$. The output $Y(s)$ is fed back through the feedback element $H(s)$, whose output $C(s)$ is compared with the input $U(s)$ at the summing junction. The difference, $E(s) = U(s) - C(s)$, is known as the error signal. Because of the negative sign associated with $C(s)$ at the summing junction, the configuration in [Figure 4.13](#) is known as a negative-feedback system.

Note that $Y(s) = E(s)G(s)$ is the input to the feedback element $H(s)$, thus $C(s) = H(s)Y(s) = H(s)E(s)G(s)$. Based on this, we define

$$\text{open-loop transfer function} = \frac{C(s)}{E(s)} = G(s)H(s)$$

Another important transfer function in the closed-loop system in [Figure 4.13](#) is defined as

$$\text{feedforward transfer function} = \frac{Y(s)}{E(s)} = G(s)$$

4.5.1.5.1 Closed-Loop Transfer Function

The closed-loop transfer function (CLTF) provides the direct relation between the (overall) input $U(s)$ and the (overall) output $Y(s)$ and is determined as follows. Referring to [Figure 4.13](#),

$$Y(s) = G(s)E(s) \stackrel{E(s)=U(s)-C(s)}{=} G(s)[U(s)-C(s)] \stackrel{C(s)=H(s)Y(s)}{=} G(s)[U(s)-H(s)Y(s)]$$

The above is then manipulated as

$$[1 + G(s)H(s)]Y(s) = G(s)U(s)$$

Finally, the CLTF is formed as

$$\text{Negative-feedback CLTF} \quad \frac{Y(s)}{U(s)} = \frac{G(s)}{1 + G(s)H(s)} \quad (4.22)$$

Similarly, in the case of a positive feedback, it can be easily shown that

$$\text{Positive-feedback CLTF} \quad \frac{Y(s)}{U(s)} = \frac{G(s)}{1 - G(s)H(s)} \quad (4.23)$$

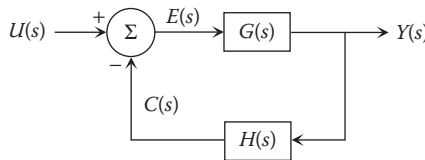


FIGURE 4.13
A negative-feedback system.

Example 4.20: Negative Feedback

In the negative-feedback system shown in [Figure 4.13](#), assume that

$$G(s) = \frac{1}{2s^2 + s + 1}, \quad H(s) = 2s + 1$$

- a. Find the CLTF.
- b.  Confirm the result of (a) in MATLAB by using `feedback`.

Solution

- a. By Equation 4.22

$$\frac{Y(s)}{U(s)} = \frac{G(s)}{1 + G(s)H(s)} = \frac{\frac{1}{2s^2 + s + 1}}{1 + \frac{2s + 1}{2s^2 + s + 1}} = \frac{1}{2s^2 + 3s + 2}$$

- b. 

```
>> sysG = tf([1],[2 1 1]); sysH = tf([2 1],[1]);
>> sysEq = feedback(sysG,sysH)
sysEq =
```

```
      1
-----
2 s^2 + 3 s + 2
```

Continuous-time transfer function.

4.5.2 Block-Diagram Reduction Techniques

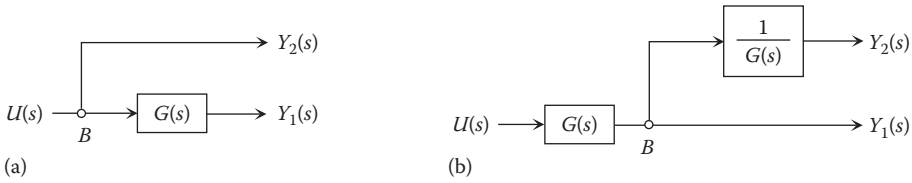
As dynamic systems get more complex in nature, so do their block-diagram representations. In these situations, the block diagram can potentially contain several summing junctions, blocks in series or parallel connections, and positive and negative feedback loops. There are a few basic rules that facilitate the process of simplifying a block diagram. These include, among others, moving a branch point and moving a summing junction, as explained in the following.

4.5.2.1 Moving a Branch Point

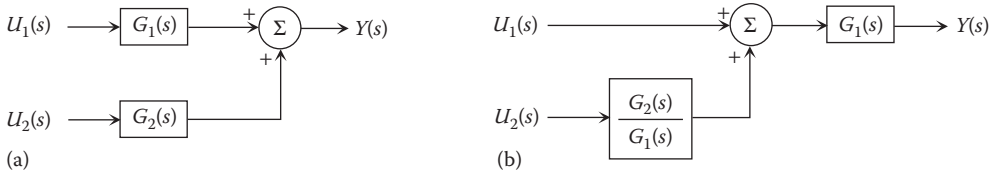
Consider the branch point B in [Figure 4.14a](#), located to the left of the block $G(s)$. The branch point may be moved to the right side of the block $G(s)$, as demonstrated in [Figure 4.14b](#). The key is for signals $Y_1(s)$ and $Y_2(s)$ to carry the same information before and after B is moved. It is readily seen that $Y_1(s) = G(s)U(s)$ in both arrangements. In addition, $Y_2(s) = U(s)$ before B is moved, and $Y_2(s) = [1/G(s)]G(s)U(s) = U(s)$ after the move.

4.5.2.2 Moving a Summing Junction

Consider the summing junction in the configuration shown in [Figure 4.15a](#). The summing junction may be moved to the left side of the block $G_1(s)$, as demonstrated in [Figure 4.15b](#).

**FIGURE 4.14**

(a) Branch point and (b) branch point moved.

**FIGURE 4.15**

(a) Summing junction and (b) summing junction moved.

Once again, the key is for the signal $Y(s)$ to carry the same information before and after the summing junction is moved. In [Figure 4.15a](#), it is easily seen that

$$Y(s) = G_1(s)U_1(s) + G_2(s)U_2(s)$$

In [Figure 4.15b](#), we have

$$Y(s) = G_1(s) \left[U_1(s) + \frac{G_2(s)}{G_1(s)} U_2(s) \right] = G_1(s)U_1(s) + G_2(s)U_2(s)$$

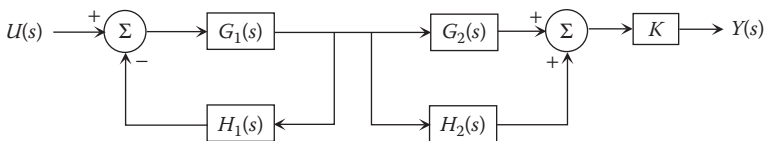
This validates the equivalence of the two arrangements.

Example 4.21: Block-Diagram Reduction

Using reduction techniques, simplify the block diagram shown in [Figure 4.16](#) to a single block with input $U(s)$ and output $Y(s)$ and subsequently find the transfer function $Y(s)/U(s)$.

Solution

There are several ways in which this block diagram can be reduced. However, the result is independent of the choices of reduction techniques and the order in which they are used. We will simplify it as follows. The block diagram is composed of a negative

**FIGURE 4.16**

Block diagram in Example 4.21.

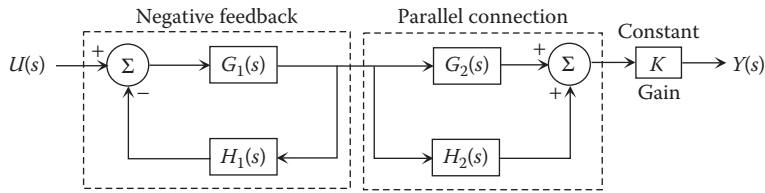


FIGURE 4.17
Feedback and parallel connection identified.

feedback, a parallel connection, and a constant gain block, as shown in Figure 4.17. The negative feedback is replaced with a single block with transfer function

$$\frac{G_1(s)}{1 + G_1(s)H_1(s)}$$

The parallel connection consists of blocks $G_2(s)$ and $H_2(s)$ and hence is replaced with a single block

$$G_2(s) + H_2(s)$$

These two single blocks are shown in Figure 4.18. The series connection in Figure 4.18 is next replaced with a single block whose transfer function is the product of the individual block transfer functions, that is,

$$\frac{G_1(s)}{1 + G_1(s)H_1(s)} [G_2(s) + H_2(s)] K$$

This yields Figure 4.19, from which the overall transfer function is easily found as

$$\frac{Y(s)}{U(s)} = \frac{KG_1(G_2 + H_2)}{1 + G_1H_1}$$

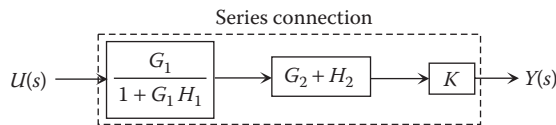


FIGURE 4.18
Feedback loop and parallel connection replaced.

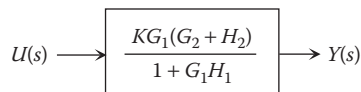


FIGURE 4.19
Simplified block diagram in Example 4.21.

4.5.2.3 Mason's Rule

So far, we have learned that when a block diagram contains several loops, each loop can be replaced with a single block with a transfer function, given by Equations 4.22 and Equation 4.23 for negative and positive feedbacks, respectively. This, in conjunction with other tactics mentioned earlier, can then help us find the overall transfer function for the block diagram. However, an alternative approach is to employ Mason's rule. We first define a forward path as the one that originates from the overall input, leading to the overall output, never moving in the opposite direction. A loop path (or a loop) is the one that originates from a certain variable and returns to the same variable. The gain of a forward path or a loop path is the product of the gains of the individual blocks that constitute the path.

4.5.2.3.1 Mason's Rule: Special Case

Suppose that all forward paths and loops in a block diagram are coupled; that is, all of them have a common segment. Then,

$$\text{overall transfer function} = \frac{\sum \text{forward path gains}}{1 - \sum \text{loop gains}} \quad (4.24)$$

Example 4.22: Mason's Rule (Special Case)

Using Mason's rule, determine the overall transfer function $Y(s)/U(s)$ for the block diagram in Figure 4.20.

Solution

Path segments have been assigned numbers for easier identification. There are two forward paths and two loops:

Forward Path	Gain	Loop	Gain
12356	G_1G_2	2372	$-G_1H_1$
12346	G_1G_3	23582	$G_1G_2H_2$

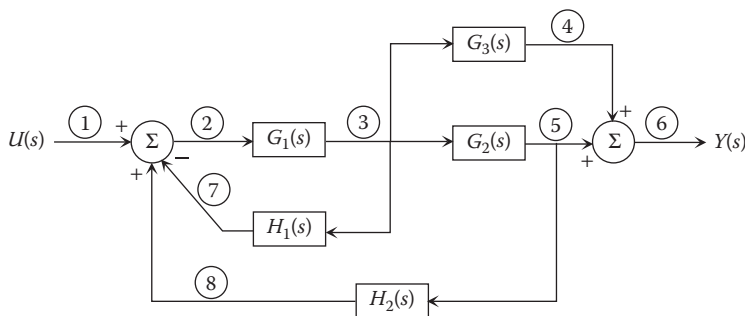


FIGURE 4.20
Block diagram in Example 4.22.

Note that the loop labeled 2372 is a negative feedback; thus, its gain is negative. Since all forward paths and loops have a common segment, labeled 23, special case of Mason's rule applies, and the overall transfer function is found via Equation 4.24, as

$$\frac{Y(s)}{U(s)} = \frac{G_1G_2 + G_1G_3}{1 + G_1H_1 - G_1G_2H_2}$$

4.5.2.3.2 Mason's Rule: General Case

In general, when all forward paths and loops are not coupled, the overall transfer function is obtained as

$$\text{overall transfer function} = \frac{\sum_{k=1}^m F_k D_k}{D}, \quad m = \text{number of forward paths} \quad (4.25)$$

where

F_k = gain of the k th forward path

$D = 1 - \sum \text{single-loop gains} + \sum \text{gain products of all non-touching two-loops}$
 $- \sum \text{gain products of all non-touching three-loops} + \dots$

D_k = same as D when the block diagram is restricted to the portion not touching the k th forward path

Example 4.23: Mason's Rule (General Case)

Using Mason's rule, find the overall transfer function $Y(s)/U(s)$ for the block diagram in Figure 4.21.

Solution

We note that not all forward paths are coupled; in particular, the two negative-feedback loops do not share a common segment. Therefore, the overall transfer function must be found via the general Mason's rule, Equation 4.25. Two forward paths and two loops are identified:

Forward Path	Gain	Loop	Gain
1234568	$G_1G_2G_4$	23(10)2	$-G_1H_1$
12378	G_1G_3	4594	$-G_2H_2$

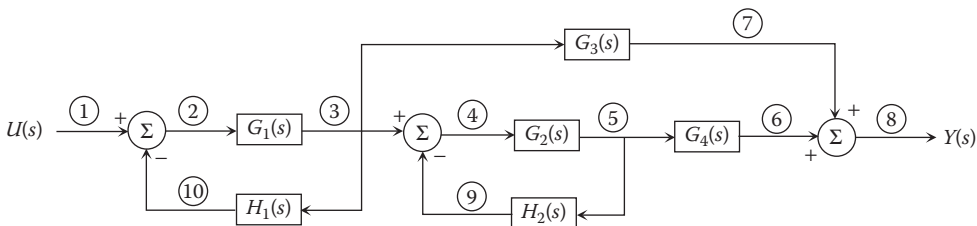


FIGURE 4.21
Block diagram in Example 4.23.

The quantities in Equation 4.25 are calculated as follows:

$$F_1 = G_1 G_2 G_4, \quad F_2 = G_1 G_3$$

$$D = 1 + \underbrace{G_1 H_1 + G_2 H_2}_{\text{Single-loop gains}} + \underbrace{G_1 H_1 G_2 H_2}_{\text{Gain product of non-touching two-loops}}$$

To find D_1 , consider the portion of the block diagram that does not touch the first forward path, 1234568. Since there are no forward paths or loops in the restricted segment, we conclude that

$$D_1 = 1$$

To determine D_2 , consider the portion of the block diagram that does not touch the second forward path, 12378. Since the restricted section contains only one single loop, 4594, we have

$$D_2 = 1 + G_2 H_2$$

Finally, Equation 4.25 yields

$$\frac{Y(s)}{U(s)} = \frac{G_1 G_2 G_4 + G_1 G_3 (1 + G_2 H_2)}{1 + G_1 H_1 + G_2 H_2 + G_1 H_1 G_2 H_2}$$

4.5.3 Block Diagram Construction from System Model

Block diagrams disclose many characteristics of dynamic systems that may not be observable by their mathematical models, such as the interrelation between the different components and variables. In what follows, we will learn how to construct partial blocks corresponding to specific segments of a system model and subsequently assemble them properly to generate the complete block diagram. Based on the block diagram, a model can then be constructed in Simulink for analysis and simulation purposes. This process will be systematically used in the subsequent chapters.

Example 4.24: Block Diagram from System Model

A dynamic system with input u and output x is governed by

$$2\ddot{x} + \dot{x} + 4x = u(t)$$

- Derive the state-variable equations and the output equation.
- Construct a block diagram by using the information in (a).
-  Build a Simulink model based on the block diagram in (b).

Solution

- The state variables are selected as $x_1 = x$ and $x_2 = \dot{x}$. Noting that the output is $y = x = x_1$, the state-variable equations and the output equation are derived as

$$\begin{cases} \dot{x}_1 = x_2 \\ \dot{x}_2 = \frac{1}{2}(-4x_1 - x_2 + u) \end{cases}, \quad y = x_1$$

b. Taking the Laplace transform of the equations in (a), we have

$$\begin{cases} sX_1(s) = X_2(s) \\ sX_2(s) = \frac{1}{2}[-4X_1(s) - X_2(s) + U(s)] \end{cases} \quad Y(s) = X_1(s)$$

The overall input $U(s)$ must appear on the far left of the diagram, and the overall output $X_1(s)$ must appear on the far right. Block diagram is built from left to right; hence, we need to start with an equation that contains the overall input $U(s)$. Thus, we start with $sX_2(s) = \frac{1}{2}[-4X_1(s) - X_2(s) + U(s)]$, which may be constructed by using a summing junction. The inputs to the summing junction are $U(s)$ with a positive sign, $4X_1(s)$ with a negative sign, and $X_2(s)$ with a negative sign. Then, the output of the summing junction is multiplied by $\frac{1}{2}$ to generate $sX_2(s)$; Figure 4.22a. This output subsequently goes through a block of $1/s$, an integrator, to generate $X_2(s)$. However, $X_2(s) = sX_1(s)$. Therefore, $X_2(s)$ will go through yet another block of $1/s$ to produce $X_1(s)$, as shown in Figure 4.22b. Since $X_1(s)$ is the overall output, we can complete the diagram, as depicted in Figure 4.22c.

c. Figure 4.23 shows a Simulink model based on the block diagram in Figure 4.22c. The input is represented by a Signal Generator (Sources library). The summing junction is represented by the Add block (Math Operations library). Double-clicking the Add block allows us to dictate the signs associated with the incoming signals. The Gain blocks and the Integrator blocks are imported from Commonly Used Blocks library. The output is stored in Out1 (Commonly Used Blocks library), as well as in Scope for simulation purposes. Using this model, the output $x(t)$ corresponding to a specified input $u(t)$ is readily generated in MATLAB.

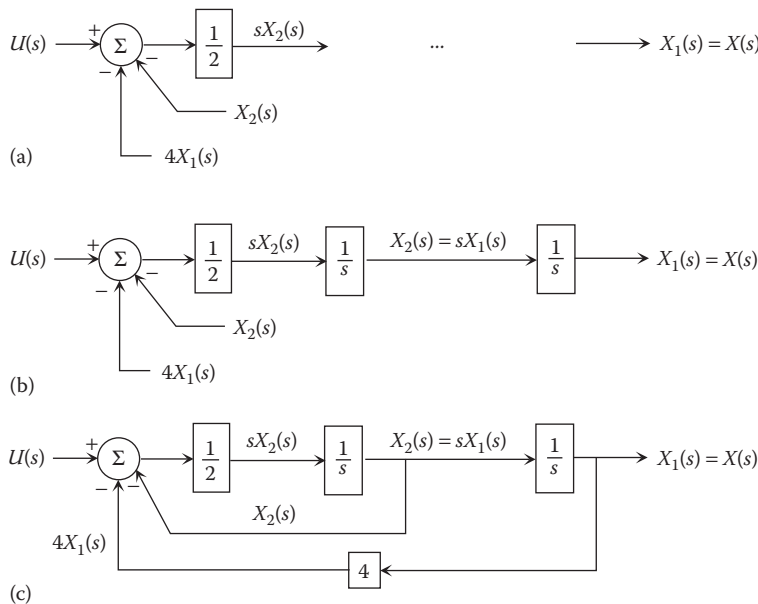


FIGURE 4.22

(a) Block diagram initiated, (b) partial block, and (c) diagram completed.

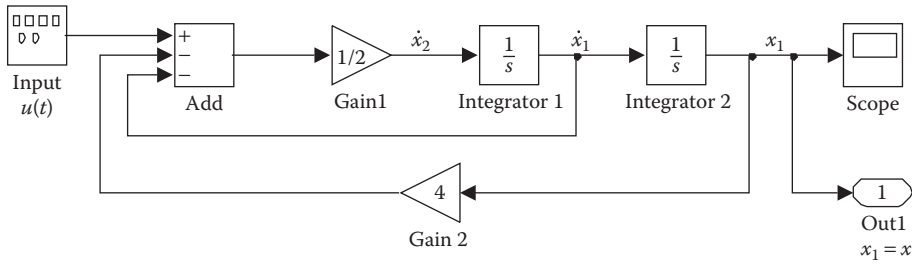


FIGURE 4.23
Simulink model in Example 4.24.

4.5.3.1 State-Space Block in Simulink

As we learned in [Section 4.2](#), linear system models can be expressed in the state-space form. The state-space form of a model can be represented in Simulink by the State-Space block imported from the Continuous library. Double-clicking the State-Space block allows us to input the state-space matrices (A , B , C , D) and the Initial Conditions.

Example 4.25: State-Space Block

Consider the dynamic system model in Example 4.24, whose state-space form is

$$\begin{cases} \dot{x} = Ax + Bu \\ y = Cx + Du \end{cases}$$

where

$$x = \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix}, A = \begin{bmatrix} 0 & 1 \\ -2 & -\frac{1}{2} \end{bmatrix}, B = \begin{bmatrix} 0 \\ \frac{1}{2} \end{bmatrix}, C = \begin{bmatrix} 1 & 0 \end{bmatrix}, D = 0, u = u(t)$$

[Figure 4.24a](#) shows the corresponding Simulink model, where there is a single output x_1 . The initial conditions can be entered by double-clicking the State-Space block. If both state variables x_1 and x_2 are the outputs, then the output equation is modified to

$$y = Cx + Du, y = \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix}, C = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}, D = \begin{bmatrix} 0 \\ 0 \end{bmatrix}, u = u(t)$$

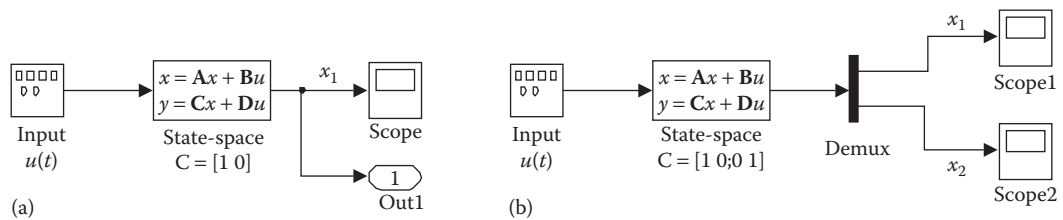


FIGURE 4.24
Simulink model in Example 4.25 with (a) single output and (b) two outputs.

In this case, enter the C matrix as the 2×2 identity matrix and import a Demux block (Signal Routing library) to split the vector signal $\mathbf{y}$ into x_1 and x_2 . This is depicted in Figure 4.24b.

PROBLEM SET 4.5

1. Reduce the block diagram in Figure 4.25 by moving the constant gain block K to the right of the summing junction. Subsequently, find the transfer function $Y(s)/U(s)$.
2. Consider the block diagram in Figure 4.26. Use the block-diagram reduction steps listed in the following to find $Y(s)/U(s)$.
 - a. Replace the single loop containing $G_1(s)$ and $H(s)$ with a single equivalent block.
 - b. Move the gain block K to the right of the summing junction.
 - c. Replace the resulting loop with a single equivalent block.
3. For the block diagram in Figure 4.27, find the transfer function $Y(s)/U(s)$ by using
 - a. Block-diagram reduction techniques.
 - b. Mason's rule.

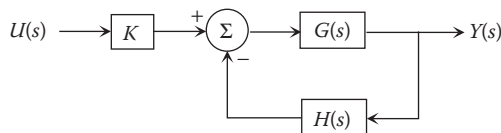


FIGURE 4.25
Problem 1.

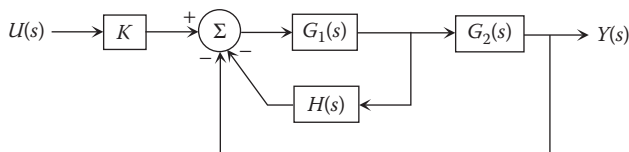


FIGURE 4.26
Problem 2.

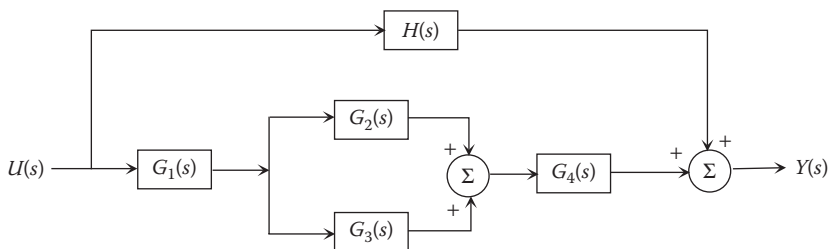


FIGURE 4.27
Problem 3.

4. Using Mason's rule, find $Y(s)/U(s)$ for the block diagram in Figure 4.16, Example 4.21.
5. Find $Y(s)/U(s)$ in Figure 4.28 by using Mason's rule.
6. Use block-diagram reduction techniques listed in the following to find the overall transfer function for the diagram in Figure 4.28.
 - a. Move the summing junction of the negative-feedback loop containing $H_2(s)$ outside of the positive-feedback loop containing $H_1(s)$.
 - b. In the ensuing diagram, replace the loop containing $H_1(s)$ with a single block.
 - c. Similarly, replace the two remaining loops with single equivalent blocks to obtain one single block whose input is $U(s)$ and output is $Y(s)$.
7. In Figure 4.29, find the transfer function $Y(s)/U(s)$ by using
 - a. Block-diagram reduction steps.
 - b. Mason's rule.
8. Find the overall transfer function in Figure 4.30 by using Mason's rule.
9. For the block diagram in Figure 4.31, find $Y(s)/U(s)$ by using
 - a. Block-diagram reduction.
 - b. Mason's rule.
10. Consider the block diagram in Figure 4.32, where $V(s)$ represents an external disturbance.
 - a. Find the transfer function $Y(s)/U(s)$ by setting $V(s) = 0$.
 - b. Find $Y(s)/V(s)$ by setting $U(s) = 0$.

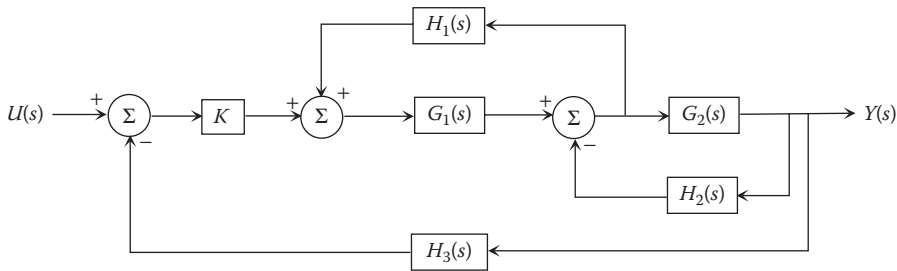


FIGURE 4.28
Problems 5 and 6.

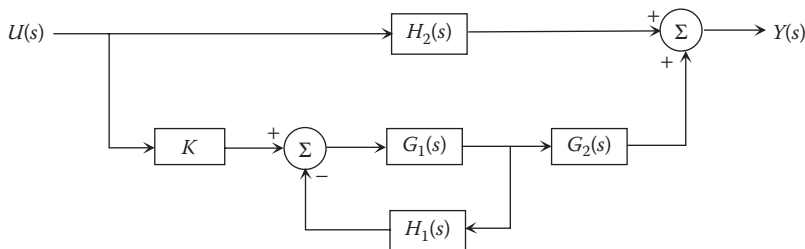


FIGURE 4.29
Problem 7.

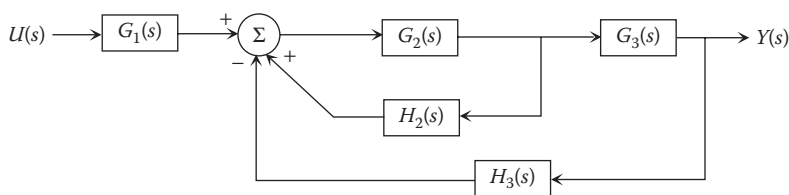


FIGURE 4.30
Problem 8.

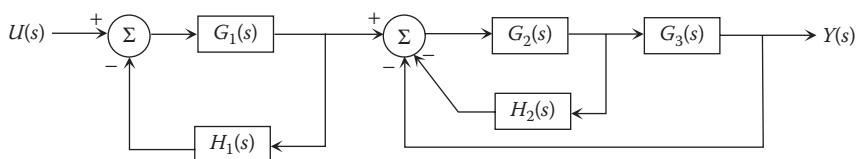


FIGURE 4.31
Problem 9.

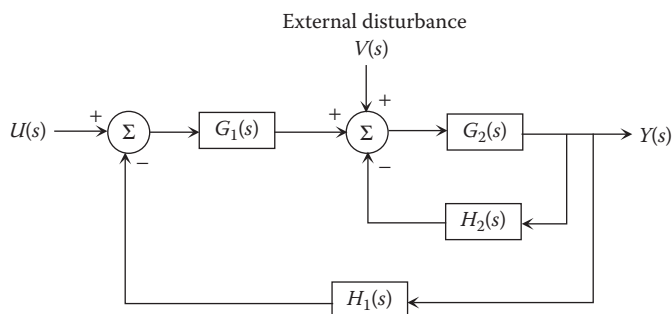


FIGURE 4.32
Problem 10.

11. The state-variable equations and the output equation of a system are given as

$$\begin{cases} \dot{x}_1 = x_2 \\ \dot{x}_2 = -\frac{1}{2}x_1 - x_2 + 2u' \end{cases} \quad y = 2x_1 + x_2$$

where u and y are the input and output, respectively.

- Construct the block diagram.
- Find the transfer function $Y(s)/U(s)$ directly from the block diagram.
- Find the transfer function $Y(s)/U(s)$ from the state-space form and compare with (b).

12. The state-space representation of a system model is given as

$$\begin{cases} \dot{\mathbf{x}} = \mathbf{Ax} + \mathbf{Bu} \\ y = \mathbf{Cx} + Du \end{cases}$$

with

$$\mathbf{x} = \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix}, \quad \mathbf{A} = \begin{bmatrix} 0 & 1 \\ -2 & -\frac{2}{3} \end{bmatrix}, \quad \mathbf{B} = \begin{bmatrix} 0 \\ 3 \end{bmatrix}, \quad \mathbf{C} = \begin{bmatrix} 1 & \frac{1}{2} \end{bmatrix}, \quad D = 0, \quad u = u$$

- Construct the block diagram.
- Find the transfer function $Y(s)/U(s)$ directly from the block diagram.
- Find the transfer function $Y(s)/U(s)$ from the state-space form and compare with (b).

In problems 13 through 16, the I/O equation of a dynamic system is provided.

- Find the state-space model.
 - Construct the block diagram based on the information in (a).
 - Determine the overall transfer function directly from the block diagram in (b).
- $3\ddot{y} + \dot{y} + 2y = \frac{1}{3}\dot{u}$
 - $\frac{2}{3}\ddot{y} + 2\dot{y} + y = \frac{1}{3}\dot{u} + 2u$
 - $2\ddot{y} + \dot{y} + y = \ddot{u} + 2\dot{u}$
 - $\ddot{y} + \frac{1}{2}\dot{y} + 2y = \frac{1}{2}\ddot{u} + u$
 - A system is described by its transfer function

$$\frac{Y(s)}{U(s)} = \frac{s + \frac{1}{3}}{s(2s^2 + s + 2)}$$

- Find the I/O equation.
 - Find the state-space model from the I/O equation.
 - Build a block diagram based on the information in (b). Find the transfer function directly from the block diagram and compare with the given transfer function.
- Repeat Problem 17 for

$$\frac{Y(s)}{U(s)} = \frac{s^2 + \frac{1}{3}}{2s^2 + s + 2}$$

- The block diagram representation of a system model is presented in Figure 4.33, where x_1 and x_2 denote the two state variables.

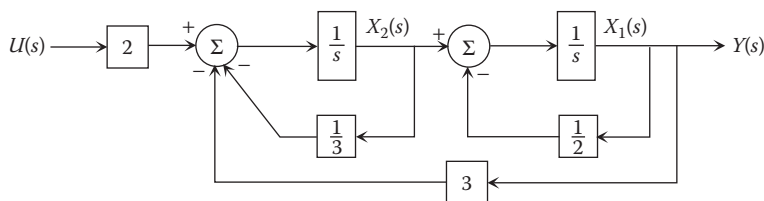
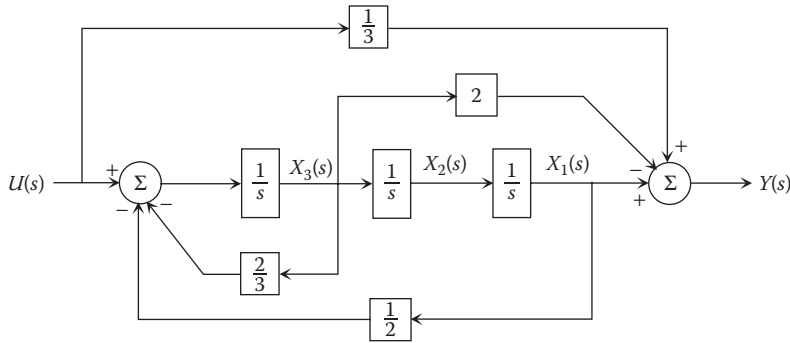


FIGURE 4.33
Problem 19.

**FIGURE 4.34**

Problem 20.

- a. Derive the state-space form directly from the block diagram.
 - b. Find the transfer function directly from the block diagram.
20. Repeat Problem 19 for the block diagram in [Figure 4.34](#), where x_1 , x_2 , and x_3 denote the state variables.

4.6 Linearization

The systems that we have studied so far have primarily been assumed linear, making their analysis somewhat straightforward. However, many dynamic systems contain elements that are inherently nonlinear, which cannot be treated as linear, except for a restricted range of operating conditions. In this section, we present a methodical approach to derive a linear approximation of a nonlinear model. We will also learn how to linearize nonlinear elements in Simulink. In [Section 8.6](#), we will investigate the response of the linearized model in relation to that of the nonlinear model.

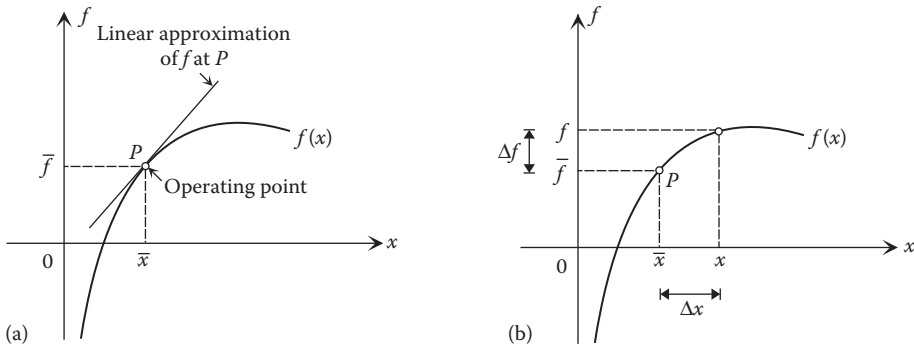
4.6.1 Linearization of a Nonlinear Element

Consider a nonlinear function f of a single variable x , as shown in [Figure 4.35a](#). The linearization of $f(x)$ will be done at a specific point $P : (\bar{x}, \bar{f})$, known as an operating point. For now, we assume that the (constant) values of $\bar{x}$ and $\bar{f}$ are available, but we will learn shortly how to find an operating point for a given system.

The values $\bar{x}$ and $\bar{f} = f(\bar{x})$ are called the nominal values of x and f , respectively. As shown in [Figure 4.35b](#), the coordinates of any point (x, f) on the graph of $f(x)$ can be expressed as

$$x(t) = \bar{x} + \Delta x(t), \quad f(t) = \bar{f} + \Delta f(t) \quad (4.26)$$

where the time-varying quantities $\Delta x(t)$ and $\Delta f(t)$ are the incremental variables for x and f , respectively. Graphically, the linear approximation of $f(x)$ is provided by the tangent line at the operating point P , with a reasonably good accuracy in a small neighborhood of P , that

**FIGURE 4.35**

(a) Linearization about an operating point and (b) incremental variables.

is, small values for $\Delta x(t)$ and $\Delta f(t)$. Analytically, this is justified by writing the Taylor series expansion of $f(x)$ about the operating point, as

$$f(x) = \underbrace{f(\bar{x}) + \left. \frac{df}{dx} \right|_P (x - \bar{x})}_{\text{Linear terms}} + \frac{1}{2!} \left. \frac{d^2 f}{dx^2} \right|_P (x - \bar{x})^2 + \dots$$

Assuming $\Delta x = x - \bar{x}$ is small, the linear approximation of $f(x)$ is achieved by retaining the first two terms, while neglecting the remaining terms containing higher powers of $x - \bar{x}$:

$$f(x) \cong f(\bar{x}) + \left. \frac{df}{dx} \right|_{\bar{x}} \Delta x \quad (4.27)$$

Therefore, the closer x is to $\bar{x}$ of the operating point, the better the linear approximation will be at x . The following example demonstrates this fact.

Example 4.26: Nonlinear Function of a Single Variable

Linearize $f(x) = x|x|$ about the operating point $P: (2.5, 6.25)$. Examine the accuracy of the linear approximation for $x = 2.45$ and $x = 2.7$.

Solution

We first note that

$$f(x) = x|x| = \begin{cases} x^2 & \text{if } x \geq 0 \\ -x^2 & \text{if } x < 0 \end{cases} \quad \xRightarrow{\text{Differentiate}} \quad \frac{df}{dx} = \begin{cases} 2x & \text{if } x \geq 0 \\ -2x & \text{if } x < 0 \end{cases}$$

Since $\bar{x} = 2.5$, Equation 4.27 yields

$$f(x) \cong f(2.5) + [2x]_{x=2.5} \Delta x = 6.25 + 5\Delta x$$

At $x = 2.45$, we have $\Delta x = x - \bar{x} = 2.45 - 2.5 = -0.05$, so that

$$f(2.45) \cong 6.25 + 5(-0.05) = 6$$

Since the exact value is $2.45^2 = 6.0025$, our linear approximation has a 0.04% relative error.

At $x = 2.7$, we have $\Delta x = x - \bar{x} = 2.7 - 2.5 = 0.2$, hence

$$f(2.7) \cong 6.25 + 5(0.2) = 7.25$$

Since the exact value is $2.7^2 = 7.29$, our linear approximation has a 0.55% relative error. The percentage relative error associated with $x = 2.7$ is much larger than that for $x = 2.45$, because it is farther away from the operating point.

4.6.1.1 Functions of Two Variables

Consider a nonlinear function $f(x, y)$ of two independent variables. In this case, the operating point is represented by $(\bar{x}, \bar{y}, f)$ and the incremental variables are defined as

$$\Delta x(t) = x(t) - \bar{x}, \quad \Delta y(t) = y(t) - \bar{y}, \quad \Delta f(t) = f(t) - \bar{f}$$

Taylor series expansion about the operating point yields

$$f(x, y) = \boxed{f(\bar{x}, \bar{y}) + \frac{\partial f}{\partial x} \bigg|_{\bar{x}, \bar{y}} (x - \bar{x}) + \frac{\partial f}{\partial y} \bigg|_{\bar{x}, \bar{y}} (y - \bar{y})} \quad \text{Linear terms}$$

$$+ \frac{1}{2!} \frac{\partial^2 f}{\partial x^2} \bigg|_{\bar{x}, \bar{y}} (x - \bar{x})^2 + \frac{1}{2!} \frac{\partial^2 f}{\partial y^2} \bigg|_{\bar{x}, \bar{y}} (y - \bar{y})^2 + \frac{\partial^2 f}{\partial x \partial y} \bigg|_{\bar{x}, \bar{y}} (x - \bar{x})(y - \bar{y}) + \dots$$

Assuming that $x - \bar{x}$ and $y - \bar{y}$ are small, the linear approximation of $f(x, y)$ in a small neighborhood of the operating point is given by

$$f(x, y) \cong f(\bar{x}, \bar{y}) + \frac{\partial f}{\partial x} \bigg|_{\bar{x}, \bar{y}} \Delta x + \frac{\partial f}{\partial y} \bigg|_{\bar{x}, \bar{y}} \Delta y \quad (4.28)$$

4.6.2 Linearization of a Nonlinear Model

The model of a nonlinear system can be linearized about the system's operating point(s) systematically through a standard procedure outlined shortly.

4.6.2.1 Operating Point

We will use notations such as $\bar{x}$ and $\bar{y}$ to represent the (fixed) coordinates of an operating point. Therefore, to find an operating point, we first replace the dependent variables such as $x(t)$ with $\bar{x} = \text{const}$. It is desired that a system operates not far from its equilibrium state. For that, we set the time-varying portion of the input $u(t)$ equal to zero. The result is an algebraic equation (for one-dimensional cases) or a system of equations (for higher dimensions) that can be solved for variables such as $\bar{x}$ and $\bar{y}$. Note that a system may have more than one operating point.

Example 4.27: Operating Point

Find the operating point(s) for a nonlinear system whose model is described by

$$2\ddot{x} + \dot{x} + x|x| = \frac{1}{4} + e^{-t}$$

Solution

Note that the input here is $\frac{1}{4} + e^{-t}$. Since $x(t)$ is the only dependent variable, the operating point is identified by $\bar{x}$. For the purpose of finding the operating point, we set the time-varying portion of the input, which is e^{-t} , equal to zero. To find the operating point, we replace x with $\bar{x}$ and the input with $\frac{1}{4}$ in the given system model. Since $\bar{x} = \text{const}$, we have $\dot{\bar{x}} = 0$ and $\ddot{\bar{x}} = 0$. Therefore,

$$\bar{x}|\bar{x}| = \frac{1}{4} \quad (a)$$

We solve this equation as follows:

Case (1): If $\bar{x} > 0$, then $|\bar{x}| = \bar{x}$ and Equation (a) simplifies to $\bar{x}^2 = \frac{1}{4}$. This has two solutions $\bar{x} = \pm \frac{1}{2}$. However, the assumption in this case is $\bar{x} > 0$, so that only the positive solution is acceptable, and

$$\bar{x} = \frac{1}{2}$$

Case (2): If $\bar{x} < 0$, then $|\bar{x}| = -\bar{x}$ and Equation (a) becomes $-\bar{x}^2 = \frac{1}{4}$, which has no real solution. Therefore, the only valid operating point corresponds to $\bar{x} = \frac{1}{2}$.

4.6.2.2 Linearization Procedure

A nonlinear system model is linearized about an operating point, as follows:

1. Find the operating point, as explained earlier.
2. Linearize the nonlinear term(s) about the operating point by means of Taylor series expansions—Equation 4.27 for functions of a single variable and Equation 4.28 for two variables.
3. In the original nonlinear model, replace variables such as x with $\bar{x} + \Delta x$ and nonlinear terms with their linear approximations of Step 2. Include the time-varying portions of the input that were previously set to zero to calculate the operating point in Step 1. The resulting system is linear in the incremental variables such as Δx .
4. Finally, use the initial conditions of the original model to calculate those for the linearized model. For instance, knowing $x(0)$ in the original system, find $\Delta x(0)$ by noting that $\Delta x(t) = x(t) - \bar{x}$, so that $\Delta x(0) = x(0) - \bar{x}$.

Example 4.28: Linearization Procedure

A nonlinear system model is given as

$$2\ddot{x} + \dot{x} + x|x| = \frac{1}{4} + e^{-t}, \quad x(0) = 0, \quad \dot{x}(0) = \frac{1}{2}$$

Derive the linearized model.

Solution

We will follow the procedure outlined earlier.

1. The operating point was determined in Example 4.27, as $\bar{x} = \frac{1}{2}$.
2. The only nonlinear element in the model is $f(x) = x|x|$, which is linearized about the operating point via Equation 4.27. Recalling Example 4.26, we find

$$f(x) \cong f(\tfrac{1}{2}) + [2x]_{x=\frac{1}{2}} \Delta x = \tfrac{1}{4} + \Delta x$$

3. In the original model, replace x with $\bar{x} + \Delta x = \frac{1}{2} + \Delta x$ and the nonlinear term with $\frac{1}{4} + \Delta x$. Reconsider the time-varying portion e^{-t} , which was previously set to zero, to obtain

$$2 \frac{d^2(\frac{1}{2} + \Delta x)}{dt^2} + \frac{d(\frac{1}{2} + \Delta x)}{dt} + (\frac{1}{4} + \Delta x) = \frac{1}{4} + e^{-t}$$

This simplifies to

$$2\Delta\ddot{x} + \Delta\dot{x} + \Delta x = e^{-t}$$

4. The initial conditions are adjusted as follows:

$$\Delta x(t) = x(t) - \tfrac{1}{2} \quad \xRightarrow{\text{Differentiate}} \quad \Delta \dot{x}(t) = \dot{x}(t) \quad \Rightarrow \quad \begin{aligned} \Delta x(0) &= x(0) - \tfrac{1}{2} = -\tfrac{1}{2} \\ \Delta \dot{x}(0) &= \dot{x}(0) = \tfrac{1}{2} \end{aligned}$$

In summary, the linearized model is derived as

$$2\Delta\ddot{x} + \Delta\dot{x} + \Delta x = e^{-t}, \quad \Delta x(0) = -\tfrac{1}{2}, \quad \Delta \dot{x}(0) = \tfrac{1}{2}$$

This is a second-order initial-value problem with constant coefficients and is easily solved to generate $\Delta x(t)$. It is important to note that $\Delta x(t)$ is not compatible with the solution $x(t)$ of the original nonlinear system. Recalling that $x(t) = \frac{1}{2} + \Delta x(t)$ in this problem, the solution $\Delta x(t)$ of the linearized model needs to be raised by $\frac{1}{2}$ unit to be compatible with $x(t)$. Figure 4.36 shows that the solutions of the nonlinear model and the linearized model indeed agree near the operating point $\bar{x} = \frac{1}{2}$. We will elaborate on this and many other related issues in Section 8.6.

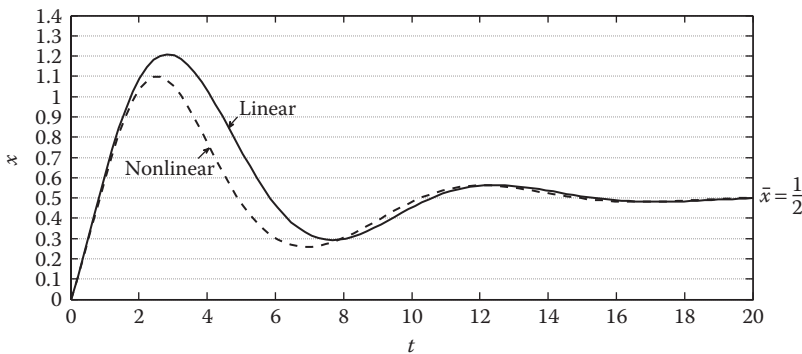


FIGURE 4.36
Solutions of a nonlinear model and its linearized form.

Example 4.29: Linearization

A system is governed by its nonlinear state-variable equations, as

$$\begin{cases} \dot{x}_1 = x_2, & x_1(0) = 0 \\ \dot{x}_2 = x_1(x_2 - 2)^2 - 4, & x_2(0) = -1 \end{cases}$$

Derive the linearized model.

Solution

We will follow the standard procedure outlined earlier. Note that the only input is -4 in the second equation.

1. Replace x_1 and x_2 with $\bar{x}_1$ and $\bar{x}_2$, respectively. Since there is no time-varying portion in the input, no further modification is needed.

$$\begin{cases} 0 = \bar{x}_2 \\ 0 = \bar{x}_1(\bar{x}_2 - 2)^2 - 4 \end{cases} \Rightarrow \begin{cases} \bar{x}_2 = 0 \\ \bar{x}_1 = 1 \end{cases}$$

Therefore, the operating point is $(\bar{x}_1, \bar{x}_2) = (1, 0)$.

2. The only nonlinear element is $f(x_1, x_2) = x_1(x_2 - 2)^2$, which will be linearized about the operating point $(1, 0)$. By Equation 4.28,

$$\begin{aligned} f(x_1, x_2) &= f(1, 0) + \left. \frac{\partial f}{\partial x_1} \right|_{(1,0)} \Delta x_1 + \left. \frac{\partial f}{\partial x_2} \right|_{(1,0)} \Delta x_2 \\ &= 4 + \left[(x_2 - 2)^2 \right]_{(1,0)} \Delta x_1 + \left[2x_1(x_2 - 2) \right]_{(1,0)} \Delta x_2 \\ &= 4 + 4\Delta x_1 - 4\Delta x_2 \end{aligned}$$

3. In the original model, replace x_1 with $\bar{x}_1 + \Delta x_1 = 1 + \Delta x_1$, x_2 with $\bar{x}_2 + \Delta x_2 = \Delta x_2$, and the nonlinear term with its linear approximation, $4 + 4\Delta x_1 - 4\Delta x_2$. No other adjustments need to be made, as no time-varying portions of input were set to zero. This yields

$$\begin{cases} \Delta \dot{x}_1 = \Delta x_2 \\ \Delta \dot{x}_2 = 4 + 4\Delta x_1 - 4\Delta x_2 - 4 \end{cases} \Rightarrow \begin{cases} \Delta \dot{x}_1 = \Delta x_2 \\ \Delta \dot{x}_2 = 4\Delta x_1 - 4\Delta x_2 \end{cases}$$

4. The initial conditions are modified as

$$\begin{aligned} \Delta x_1(0) &= x_1(0) - \bar{x}_1 & \Rightarrow & \Delta x_1(0) = -1 \\ \Delta x_2(0) &= x_2(0) - \bar{x}_2 & \Rightarrow & \Delta x_2(0) = -1 \end{aligned}$$

Therefore, the linearized model is

$$\begin{cases} \Delta \dot{x}_1 = \Delta x_2, & \Delta x_1(0) = -1 \\ \Delta \dot{x}_2 = 4\Delta x_1 - 4\Delta x_2, & \Delta x_2(0) = -1 \end{cases}$$

This linear system is then solved for $\Delta x_1(t)$ and $\Delta x_2(t)$. Once again, it must be noted that $\Delta x_1(t)$ and $\Delta x_2(t)$ are not compatible with the coordinates $x_1(t)$ and $x_2(t)$ of the original nonlinear system. Compatibility is achieved by taking into account that $x_1 = 1 + \Delta x_1$ and $x_2 = \Delta x_2$.

4.6.2.3 Small-Angle Linearization

Mathematical models of some dynamic systems contain nonlinear elements that involve trigonometric functions of an angle. One particular area where such cases occur is the rotational motion of a mechanical system; see [Section 5.3](#). These types of models can be linearized as long as the angle remains small. In particular, if $\theta \ll 1$ radian, then

$$\begin{aligned}\sin \theta &\cong \theta \\ \cos \theta &\cong 1 \\ \theta \dot{\theta}^2 &\cong 0\end{aligned}\tag{4.29}$$

Example 4.30: Small-Angle Linearization

The governing equations for a dynamic system have been derived as

$$\begin{cases} 2\ddot{x} + \ddot{\theta} - \dot{\theta}^2 \sin \theta + \dot{x} + x = f(t) \\ \ddot{\theta} + \ddot{x} \cos \theta + 10 \sin \theta = 0 \end{cases}$$

where $f(t)$ is the system input.

- Derive the linearized model for $\theta \ll 1$ radian.
- Assuming that x and θ are the outputs, find the state-space form for the linear model.

Solution

- Using the first two approximations in Equation 4.29, the governing equations reduce to

$$\begin{cases} 2\ddot{x} + \ddot{\theta} - \dot{\theta}^2 \theta + \dot{x} + x = f(t) \\ \ddot{\theta} + \ddot{x} + 10\theta = 0 \end{cases} \quad \begin{matrix} \dot{\theta}^2 \theta \cong 0 \\ \Rightarrow \end{matrix} \quad \begin{aligned} &\text{(a)} \begin{cases} 2\ddot{x} + \ddot{\theta} + \dot{x} + x = f(t) \\ \ddot{\theta} + \ddot{x} + 10\theta = 0 \end{cases} \\ &\text{(b)} \end{aligned}$$

- In its present form, the linearized model cannot be transformed into state-variable equations, because both $\ddot{x}$ and $\ddot{\theta}$ appear in the same equation. However, the remedy is to manipulate the two equations, labeled Equations a and b, to eliminate the unwanted variables, as follows. From Equation b, we find that $\ddot{\theta} = -10\theta - \ddot{x}$. We insert it into Equation a and simplify to obtain

$$\ddot{x} + \dot{x} + x - 10\theta = f(t)\tag{c}$$

This equation is now in the correct form. Next, from Equation b, we find that $\ddot{x} = -10\theta - \ddot{\theta}$. We insert it into Equation a and simplify to arrive at

$$\ddot{\theta} + 20\theta - \dot{x} - x = -f(t)\tag{d}$$

In summary,

$$\begin{aligned} &\text{(c)} \begin{cases} \ddot{x} + \dot{x} + x - 10\theta = f(t) \\ \ddot{\theta} + 20\theta - \dot{x} - x = -f(t) \end{cases} \\ &\text{(d)} \end{aligned}$$

Selecting the state variables as $x_1 = x$, $x_2 = \theta$, $x_3 = \dot{x}$, and $x_4 = \dot{\theta}$, the state-variable equations are subsequently formed as

$$\begin{cases} \dot{x}_1 = x_3 \\ \dot{x}_2 = x_4 \\ \dot{x}_3 = -x_3 - x_1 + 10x_2 + f(t) \\ \dot{x}_4 = -20x_2 + x_3 + x_1 - f(t) \end{cases}$$

Noting that there is only one input, $f(t)$, the state equation is written as

$$\dot{\mathbf{X}} = \mathbf{A}\mathbf{x} + \mathbf{B}u, \mathbf{X} = \begin{Bmatrix} x_1 \\ x_2 \\ x_3 \\ x_4 \end{Bmatrix}, \mathbf{A} = \begin{bmatrix} 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 1 \\ -1 & 10 & -1 & 0 \\ 1 & -20 & 1 & 0 \end{bmatrix}, \mathbf{B} = \begin{bmatrix} 0 \\ 0 \\ 1 \\ -1 \end{bmatrix}, u = f(t)$$

Since x and θ are the outputs, the output equation is formed as

$$\mathbf{y} = \mathbf{C}\mathbf{x} + \mathbf{D}u, \mathbf{y} = \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix}, \mathbf{C} = \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \end{bmatrix}, \mathbf{D} = \begin{bmatrix} 0 \\ 0 \end{bmatrix}$$

These types of systems will be discussed in greater detail in [Section 8.6](#)

4.6.3 Linearization in Simulink

The state-space form of the linearized model of a nonlinear system can be extracted in Simulink by using the 'linearize' built-in function:

`linearize` Obtains a linear model from a Simulink model.

`LIN = linearize('sys',OP,IO)` takes a Simulink model name, 'sys', an operating point object, OP, and an I/O object, IO, as inputs and returns a linear time-invariant state-space model, LIN. The operating point object is created with the function `OPERPOINT` or `FINDOP`. The linearization I/O object is created with the function `GETLINIO` or `LINIO`. Both OP and IO must be associated with the same Simulink model, sys.

The function is used in conjunction with 'findop' and 'getlinio'. The function call for 'findop' is as follows:

```
findop Find operating points from specifications or simulation
[OP_POINT,OP_REPORT] = findop('model',OP_SPEC) finds an operating point,
OP_POINT, of the model, 'model', from specifications given in OP_SPEC.
```

op_spec is an operating point specification object and is created with the function 'operspec'. Specifications on the operating points, such as minimum and maximum values, initial guesses, and known values, are specified by editing opspec. To find equilibrium (or steady-state) operating points, set the `SteadyState` property of the states and inputs in opspec to 1. The findop function uses optimization to find operating points that closely meet the specifications listed in opspec. The function call for 'getlinio' is as follows:

```
IO = getlinio('sys') finds all linearization annotations in a Simulink
model, 'sys'. The vector of objects returned from this function call
has an entry for each linearization annotation in a model.
```

To initiate the process, first build a Simulink model, 'sys', and identify a linearization input point and a linearization output point. To identify a linearization input point, right-click the desired signal and select **Linear Analysis Points > Input Perturbation**. To identify a linearization output point, right-click the desired signal and select **Linear Analysis Points > Output Measurement**.

Example 4.31: Linearization in Simulink

A nonlinear system model is described by

$$2\ddot{x} + \dot{x} + x^3 = 1 + \sin 2t, \quad x(0) = 0, \quad \dot{x}(0) = 1$$

In order to build a Simulink model, the nonlinear state-variable equations must first be formed. Choosing state variables $x_1 = x$ and $x_2 = \dot{x}$, we find

$$\begin{cases} \dot{x}_1 = x_2 & x_1(0) = 0 \\ \dot{x}_2 = \frac{1}{2}(-x_2 - x_1^3 + 1 + \sin 2t)' & x_2(0) = 1 \end{cases} \quad (a)$$

Following the procedure outlined in [Section 4.5](#), we build the model shown in [Figure 4.37](#) for the nonlinear system in Equation a and save it as 'Example431'. Double-clicking an integrator block allows us to input the appropriate initial condition for the output signal of that block. Since the first integrator has x_2 as its output, we use $x_2(0) = 1$. For the second integrator, $x_1(0) = 0$. The nonlinear element x_1^3 is handled by the function block from the Simulink User-Defined Functions library, where u is used as the input variable name. The linearization input point is chosen at the output signal of the summing junction after the gain block, the linearization output point at the output signal x_1 .

The linear model is extracted from the nonlinear system in Equation (a), as follows. Note that the model 'Example431' must be open before we can proceed.

```
>> sys='Example431';
>> load_system(sys);
>> opspec = operspec(sys);

% Specify the properties of the first state variable
>> opspec.States(1).SteadyState = 1;
>> opspec.States(1).x = 0;      % Initial value
>> opspec.States(1).Min = 0;    % Minimum value
```

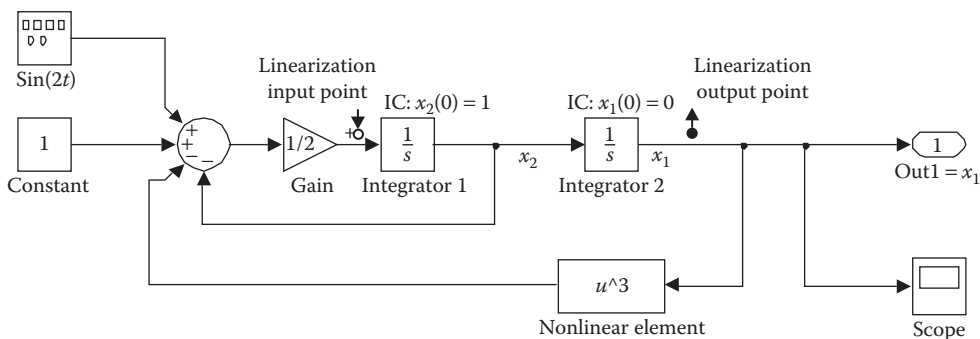


FIGURE 4.37
Simulink model in Example 4.31.

```

% Specify the properties of the second state variable
>> opspec.States(2).SteadyState = 1;
>> opspec.States(2).x = 1;      % Initial value
>> opspec.States(2).Min = 0;    % Minimum value

% Find the operating point based on the above specifications
>> [op,opreport] = findop(sys,opspec);

Operating Point Search Report:
-----
Operating Report for the Model Example431.
(Time-Varying Components Evaluated at time t=0)

Operating point specifications were successfully met.
States:
-----
(1.) Example431/Integrator 1
      x:          0      dx:          0 (0)
(2.) Example431/Integrator 2
      x:          1      dx:          0 (0)

Inputs: None
-----

Outputs:
-----
(1.) Example431/Out1 = x1
      y:          1      [-Inf Inf]

% Get the linearization annotations
>> IO=getlinio(sys);

% Extract the linear state-space model
>> LIN = linearize('Example431',op,IO)

LIN =

a =
      Integrator 1  Integrator 2
Integrator 1      -0.5      -1.5      % Controller canonical form
Integrator 2           1           0

b =
      Gain
Integrator 1      1
Integrator 2      0

c =
      Integrator 1  Integrator 2
x1              0              1

d =
      Gain
x1          0

Continuous-time state-space model.

```

NOTES

- The state matrix is in the controller canonical form.
- The input in the linearized model contains the time-varying portion of the input in the original nonlinear model. In the current example, since the input for the nonlinear system is $1 + \sin 2t$, the input for the linear model is simply $\sin 2t$.

We will validate the Simulink results as follows. The operating point satisfies $\bar{x}^3 = 1$, so that $\bar{x} = 1$. The nonlinearity $f(x) = x^3$ is linearized about $\bar{x} = 1$, as

$$f(x) = x^3 \cong f(1) + [3x^2]_{x=1} \Delta x = 1 + 3\Delta x$$

Substitute $x = 1 + \Delta x$ and the linear approximation into the original equation:

$$2(1 + \Delta x) + (1 + \Delta x) + [1 + 3\Delta x] = 1 + \sin 2t \Rightarrow 2\Delta\ddot{x} + \Delta\dot{x} + 3\Delta x = \sin 2t$$

To derive the controller canonical form, choose the state variables as $\Delta x_1 = \Delta\dot{x}$ and $\Delta x_2 = \Delta x$, and the state-variable equations for the linearized model are obtained as

$$\begin{cases} \Delta\dot{x}_1 = \frac{1}{2}(-\Delta x_1 - 3\Delta x_2 + \sin 2t) \\ \Delta\dot{x}_2 = \Delta x_1 \end{cases}$$

In vector form,

$$\begin{cases} \Delta\dot{\mathbf{x}} = \mathbf{A}\Delta\mathbf{x} + \mathbf{B}u \\ y = \mathbf{C}\Delta\mathbf{x} + Du \end{cases} \quad (\text{b})$$

where

$$\Delta\mathbf{x} = \begin{Bmatrix} \Delta x_1 \\ \Delta x_2 \end{Bmatrix}, \quad \mathbf{A} = \begin{bmatrix} -\frac{1}{2} & -\frac{3}{2} \\ 1 & 0 \end{bmatrix}, \quad \mathbf{B} = \begin{bmatrix} 1 \\ 0 \end{bmatrix}, \quad \mathbf{C} = [0 \quad 1], \quad D = 0, \quad u = \frac{1}{2}\sin 2t$$

This confirms the set of results generated in Simulink. Linear analysis of nonlinear systems will be further discussed in [Section 8.6](#).

PROBLEM SET 4.6

In Problems 1 through 10, the mathematical model of a nonlinear system is provided. Using the procedure outlined in this section, find the operating point and derive the linearized model.

1. $\ddot{x} + 2\dot{x} + x^3 = u(t)$, $u(t) = \text{unit-step}$, $x(0) = \frac{1}{2}$, $\dot{x}(0) = 0$
2. $\ddot{x} + \frac{1}{3}x^3 = \frac{1}{3} + \sin t$, $x(0) = 1$, $\dot{x}(0) = 1$
3. $\frac{1}{2}\ddot{x} + \dot{x} + 3x|x| = \frac{1}{3} + \cos t$, $x(0) = -\frac{1}{2}$, $\dot{x}(0) = 0$
4. $\ddot{x} + \dot{x} + x\sqrt{|x|} = \frac{1}{8} + \sin(\frac{1}{2}t)$, $x(0) = 0$, $\dot{x}(0) = 1$
5. $\ddot{x} + \frac{2}{3}\dot{x} + f(x) = 1 + e^{-t/3}$, $x(0) = 1$, $\dot{x}(0) = \frac{1}{3}$, $f(x) = \begin{cases} 3\sqrt{x} & \text{if } x \geq 0 \\ -3\sqrt{|x|} & \text{if } x \leq 0 \end{cases}$

$$6. \frac{1}{3}\ddot{x} + \dot{x} + f(x) = 3 + \sin t, \quad x(0) = 0, \quad \dot{x}(0) = -1, \quad f(x) = \begin{cases} \frac{1}{2}(1 - e^{-x/2}) & \text{if } x \geq 0 \\ -\frac{1}{2}(1 - e^{-x/2}) & \text{if } x < 0 \end{cases}$$

$$7. \begin{cases} \dot{x}_1 = \frac{1}{2}x_2 - x_1 & x_1(0) = \frac{3}{4} \\ \dot{x}_2 = 2x_2^{-1} + 1 + e^{-t} & x_2(0) = 1 \end{cases}$$

$$8. \begin{cases} \dot{x}_1 = x_2 & x_1(0) = 1 \\ \dot{x}_2 = -x_1 - x_1|x_1| - x_2^3 - 2 + \sin(\frac{1}{3}t) & x_2(0) = -1 \end{cases}$$

$$9. \begin{cases} \dot{x}_1 = -x_2 - x_1|x_1| - 2 + \sin 2t & x_1(0) = \frac{2}{3} \\ \dot{x}_2 = x_1 - x_2 - 2 & x_2(0) = 0 \end{cases}$$

$$10. \begin{cases} \dot{x}_1 = -x_1 + 3x_2^3 & x_1(0) = 0 \\ \dot{x}_2 = x_1 + 24 + \frac{1}{3}e^{-t/3} & x_2(0) = 1 \end{cases}$$

In Problems 11 through 14, a nonlinear model is given.

- Obtain the linear state-space form in Simulink.
- Derive the linearized model analytically to confirm the result of (a).

$$11. \ddot{x} + \dot{x} + x|x| = 1 + \sin t, \quad x(0) = 0, \quad \dot{x}(0) = 1$$

$$12. \ddot{x} + \dot{x}^3 + x = u(t), \quad u(t) = \text{unit-step}, \quad x(0) = 1, \quad \dot{x}(0) = 1$$

$$13. \ddot{x} + \dot{x}|x| + x = \frac{1}{2} + \cos(\frac{1}{2}t), \quad x(0) = \frac{1}{2}, \quad \dot{x}(0) = 0$$

$$14. 2\ddot{x} + x^3 = \frac{1}{8} + \sin t, \quad x(0) = 1, \quad \dot{x}(0) = 1$$

4.7 Summary

A set of generalized coordinates is a set of coordinates that completely describes the motion of a system. If there are n generalized coordinates $q_1, q_2, \dots, q_n$, the system model is described by the configuration form

$$\begin{cases} \ddot{q}_1 = f_1(q_1, q_2, \dots, q_n, \dot{q}_1, \dot{q}_2, \dots, \dot{q}_n, t) \\ \ddot{q}_2 = f_2(q_1, q_2, \dots, q_n, \dot{q}_1, \dot{q}_2, \dots, \dot{q}_n, t) \\ \dots \\ \ddot{q}_n = f_n(q_1, q_2, \dots, q_n, \dot{q}_1, \dot{q}_2, \dots, \dot{q}_n, t) \end{cases}$$

where $\dot{q}_1, \dot{q}_2, \dots, \dot{q}_n$ are the generalized velocities and $f_1, f_2, \dots, f_n$, known as the generalized forces, are functions of q_i and $\dot{q}_i$ ($i = 1, 2, \dots, n$) and time t .

Models of dynamic systems governed by n -dimensional systems of second-order differential equations are represented by the standard, second-order matrix form

$$\mathbf{M}\ddot{\mathbf{x}} + \mathbf{C}\dot{\mathbf{x}} + \mathbf{K}\mathbf{x} = \mathbf{f}$$

where

- $\mathbf{x}_{n \times 1}$ = configuration vector and $\mathbf{f}_{n \times 1}$ = vector of external forces
- $\mathbf{M}_{n \times n}$ = mass matrix, $\mathbf{C}_{n \times n}$ = damping matrix, and $\mathbf{K}_{n \times n}$ = stiffness matrix

State variables x_i ($i = 1, 2, \dots, n$) form the smallest set of independent variables that completely describes the state of a system. There are as many state variables for a system as there are initial conditions required to completely solve the model. The state variables are those variables for which initial conditions are needed. For a linear system with state variables $x_1, x_2, \dots, x_n$, inputs $u_1, u_2, \dots, u_m$, and outputs $y_1, y_2, \dots, y_p$, the state-space form is

$$\begin{cases} \dot{\mathbf{x}} = \mathbf{A}_{n \times n} \mathbf{x}_{n \times 1} + \mathbf{B}_{n \times m} \mathbf{u}_{m \times 1} \\ \mathbf{y}_{p \times 1} = \mathbf{C}_{p \times n} \mathbf{x}_{n \times 1} + \mathbf{D}_{p \times m} \mathbf{u}_{m \times 1} \end{cases}$$

where

- $\mathbf{x}_{n \times 1}$ = state vector, $\mathbf{A}$ = state matrix, $\mathbf{B}$ = input matrix, and $\mathbf{y}$ = output vector
- $\mathbf{C}$ = output matrix, $\mathbf{D}$ = direct transmission matrix, and $\mathbf{u}$ = input vector

If $u(t)$ is an input and $y(t)$ is an output of a system, then the I/O equation is

$$\begin{aligned} y^{(n)} + a_1 y^{(n-1)} + \dots + a_{n-1} \dot{y} + a_n y &= b_0 u^{(m)} + b_1 u^{(m-1)} \\ &+ \dots + b_{m-1} \dot{u} + b_m u, \quad m \leq n \end{aligned}$$

where $a_1, \dots, a_n$ and $b_0, b_1, \dots, b_m$ are constants.

A transfer function is the ratio of the Laplace transforms of the output and the input. If a system has q inputs and r outputs, the qr transfer functions are assembled in an $r \times q$ transfer matrix, denoted by $\mathbf{G}(s) = [G_{ij}(s)]$, where $i = 1, 2, \dots, r$ and $j = 1, 2, \dots, q$.

A block diagram is an interconnection of blocks, each block corresponding to an operation carried out by a component, such that the block diagram as a whole agrees with the system's mathematical model.

The linearization of a nonlinear system model is performed as follows:

- Find the operating point(s) by assuming that the dependent variables are constants and setting the time-varying portions of inputs equal to zero.
- Linearize the nonlinear term(s) about the operating point(s).
- In the original nonlinear model, express variables in terms of incremental variables, and replace nonlinear terms with their linear approximations. The resulting system is linear in the incremental variables.
- Find the initial conditions for the incremental variables.

The state-space linear model of a nonlinear system can be extracted in Simulink by using the 'linearize' built-in function. The function is used in conjunction with 'findop' and 'getlinio'. The inputs to the 'findop' function are the model, 'sys', and 'opspec', which is an operating point specification object, and is created with the function 'operspec'. Specifications on the operating points, such as minimum and maximum values, initial

guesses, and known values, are specified by editing 'opspec'. To find equilibrium (or steady-state) operating points, set the 'SteadyState' property of the states and inputs in 'opspec' to 1. The 'findop' function uses optimization to find operating points that closely meet the specifications listed in 'opspec'.

Review Problems

1. The mathematical model of a dynamic system is derived as

$$\begin{cases} \ddot{x}_1 + \dot{x}_1 + x_1 - 3x_2 = 0 \\ 2\dot{x}_2 - x_1 - \frac{1}{4}x_2 = f(t) \end{cases}$$

- a. If $f(t)$ is the input and x_1 and $\dot{x}_1$ are the outputs, obtain the state-space form.
 - b.  Determine if the system is stable (Problem Set 4.2) by examining the state-space form.
 - c. Find the transfer matrix directly from the given model.
 - d.  Confirm the result of (c) by using `ss2tf`.
2. A system is described by its governing equations

$$\begin{cases} \frac{1}{2}\ddot{x}_1 + x_1 - 2(x_2 - x_1) = f(t) \\ \ddot{x}_2 + 2(x_2 - x_1) = 0 \end{cases}$$

where f is the input and x_2 and $\dot{x}_2$ are the outputs.

- a. Obtain the state-space form.
 - b.  Find the transfer matrix directly from the state-space form.
 - c. Find the transfer matrix by using the governing equations, and compare with (b).
3. A dynamic system with input f and output $\dot{x}$ is modeled as

$$\frac{3}{4}\ddot{x} + \dot{x} + kx = f(t), \quad k = \text{const} > 0$$

- a. Find the state-space form, and determine the value(s) of k for which the system is stable.
 - b. Find the transfer function directly from the state-space form.
 - c. Find the transfer function directly from the given model, and compare with (b).
4. The I/O equation of a system is derived as

$$2\ddot{y} + \frac{1}{2}\dot{y} + y = \ddot{u} + \frac{1}{3}\dot{u} + u$$

- a. Find the state-space representation.
- b. Find the transfer function from the state-space form.

5. A system's transfer function is given as

$$\frac{Y(s)}{U(s)} = \frac{s^2(2s+1)}{\frac{1}{3}s^3 + s^2 + s + 2}$$

- Find the state-space form.
- Find the transfer function from (a).

6. Find the I/O equation of a SISO system whose state-space form is given as

$$\begin{cases} \dot{\mathbf{x}} = \mathbf{A}\mathbf{x} + \mathbf{B}u \\ y = \mathbf{C}\mathbf{x} + Du \end{cases}$$

where

$$\mathbf{A} = \begin{bmatrix} 0 & 1 \\ -\frac{1}{4} & -\frac{1}{2} \end{bmatrix}, \mathbf{B} = \begin{bmatrix} 0 \\ \frac{3}{2} \end{bmatrix}, u = u, \mathbf{C} = \begin{bmatrix} \frac{2}{3} & \frac{1}{3} \end{bmatrix}, D = 0$$

7. Repeat Problem 6 for

$$\mathbf{A} = \begin{bmatrix} 0 & 1 \\ -1 & -4 \end{bmatrix}, \mathbf{B} = \begin{bmatrix} 0 \\ 3 \end{bmatrix}, u = u, \mathbf{C} = \begin{bmatrix} 2 & -1 \end{bmatrix}, D = 2$$

8. Find all possible I/O equations for a system with state-space form

$$\begin{cases} \dot{\mathbf{x}} = \mathbf{A}\mathbf{x} + \mathbf{B}u \\ \mathbf{y} = \mathbf{C}\mathbf{x} + \mathbf{D}u \end{cases}$$

where

$$\mathbf{A} = \begin{bmatrix} 0 & 1 \\ -\frac{1}{2} & -1 \end{bmatrix}, \mathbf{B} = \begin{bmatrix} 0 \\ \frac{1}{3} \end{bmatrix}, u = u, \mathbf{C} = \begin{bmatrix} 1 & 0 \\ 0 & 2 \end{bmatrix}, \mathbf{D} = \begin{bmatrix} 0 \\ 0 \end{bmatrix}$$

In Problems 9 through 12, given the block diagram representation of the system, find the overall transfer function.

9. [Figure 4.38](#)

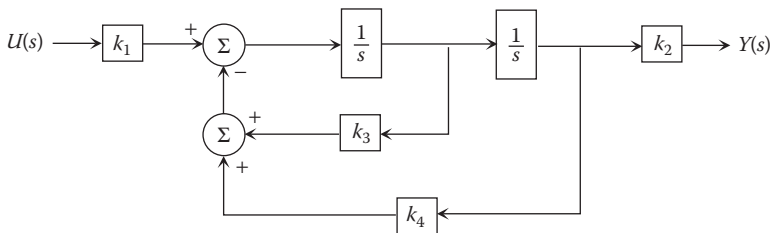


FIGURE 4.38
Problem 9.

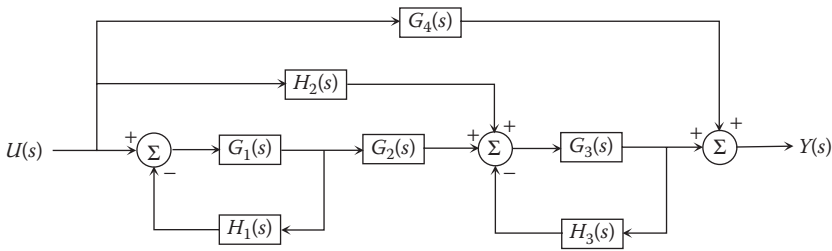


FIGURE 4.39
Problem 10.

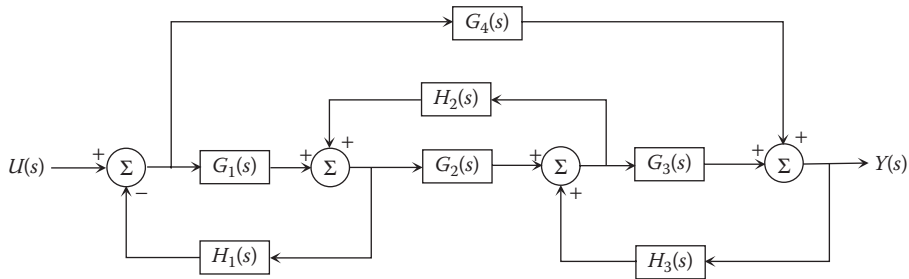


FIGURE 4.40
Problem 11.

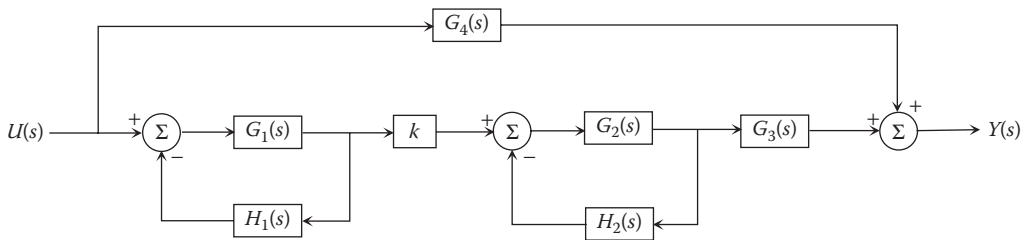


FIGURE 4.41
Problem 12.

10. [Figure 4.39](#)
11. [Figure 4.40](#)
12. [Figure 4.41](#)

In Problems 13 and 14, given the block diagram, find the system's I/O equation.

13. [Figure 4.42](#)
14. [Figure 4.43](#)

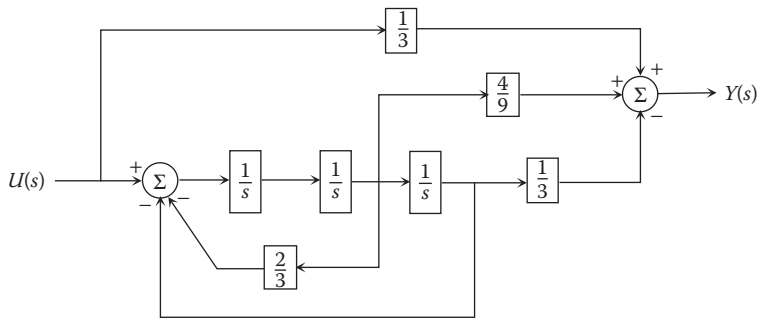


FIGURE 4.42
Problem 13.

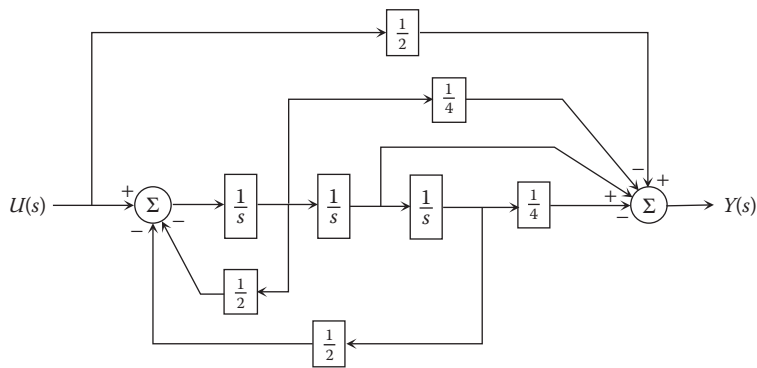


FIGURE 4.43
Problem 14.

In Problems 15 and 16, given the state-space representation of the system model, construct the appropriate block diagram, and directly use it to find the transfer matrix.

$$15. \begin{cases} \dot{\mathbf{x}} = \mathbf{A}\mathbf{x} + \mathbf{B}u \\ y = \mathbf{C}\mathbf{x} + Du \end{cases}$$

where

$$\mathbf{x} = \begin{bmatrix} x_1 \\ x_2 \\ x_3 \end{bmatrix}, \quad \mathbf{A} = \begin{bmatrix} 0 & 1 & 0 \\ 0 & 0 & 1 \\ -\frac{1}{2} & -1 & -\frac{1}{2} \end{bmatrix}, \quad \mathbf{B} = \begin{bmatrix} 0 \\ 0 \\ 1 \end{bmatrix}, \quad \mathbf{C} = [1 \quad 2 \quad 0], \quad D = 0, \quad u = u$$

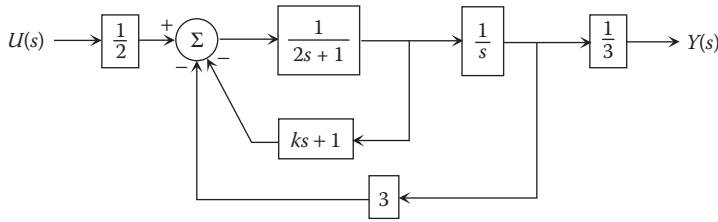


FIGURE 4.44
Problem 17.

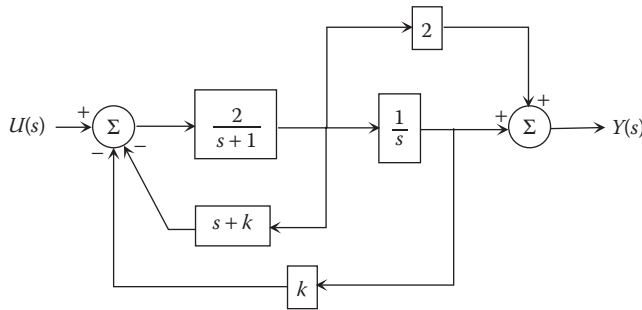


FIGURE 4.45
Problem 18.

$$16. \begin{cases} \dot{\mathbf{x}} = \mathbf{Ax} + \mathbf{Bu} \\ \mathbf{y} = \mathbf{Cx} + \mathbf{Du} \end{cases}$$

where

$$\mathbf{x} = \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix}, \quad \mathbf{A} = \begin{bmatrix} 0 & 1 \\ -1 & -3 \end{bmatrix}, \quad \mathbf{B} = \begin{bmatrix} 0 \\ 1 \end{bmatrix}, \quad \mathbf{C} = \begin{bmatrix} 1 & 0 \\ 0 & 2 \end{bmatrix}, \quad \mathbf{D} = \begin{bmatrix} 0 \\ 0 \end{bmatrix}, \quad u = u$$

17. A system is stable if the poles of the overall transfer function lie in the left half plane; these poles are the same as the eigenvalues of the state matrix. Consider the block diagram representation of a system shown in Figure 4.44, where $k > 0$ is a parameter. Determine the range of values of k for which the system is stable.
18. Repeat Problem 17 for the block diagram in Figure 4.45, where $k > 0$ is a parameter.
19. Derive the linearized model for the nonlinear system, described by

$$\begin{cases} \dot{x}_1 = x_1|x_1| + x_2 - 1 + \sin t & x_1(0) = -1 \\ \dot{x}_2 = -x_1 - x_2 - 1 & x_2(0) = 0 \end{cases}$$

20. Repeat Problem 19 for

$$\begin{cases} \dot{x}_1 = x_2 + 2\sin t & x_1(0) = 1 \\ \dot{x}_2 = x_1^3 + 3x_2 - 1 & x_2(0) = 0 \end{cases}$$

5

Mechanical Systems

This chapter discusses the modeling techniques for mechanical systems in either translational or rotational motion, or both. We begin this chapter with an introduction of mechanical elements, including mass elements, spring elements, and damper elements. The concept of equivalence is discussed, which simplifies the modeling of systems in many applications. We then review Newton's second law and apply it to translational systems. For rotational systems, moment equations are used to obtain dynamic models. For systems involving both translational and rotational motions, equations of motion can be derived using the force/moment approach, based on Newtonian mechanics, or the energy method, based on analytical mechanics. Examples are given to illustrate both methods, followed by a brief coverage of gear–train systems. The chapter concludes with simulation of mechanical systems, using MATLAB®, Simulink®, and Simscape™ computer tools.

5.1 Mechanical Elements

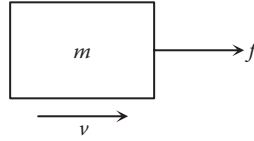
The objective of this chapter is to show how one can obtain mathematical models of mechanical systems. Because a real mechanical system is usually complicated, simplifying assumptions must be made to reduce the system to an idealized model, which consists of interconnected elements. The behavior of the mathematical model can then approximate that of the real system.

A mathematical model of a mechanical system can be constructed based on physical laws (such as Newton's laws and the conservation of energy) that the elements and their interconnections must obey. Elements can be broadly divided into three classes according to whether element forces are proportional to accelerations, proportional to displacements, or proportional to velocities. Correspondingly, they can be divided into elements that store and release kinetic energy, store and release potential energy, and dissipate energy. In this section, element equations relating the external forces to the associated element variables are presented.

5.1.1 Mass Elements

Figure 5.1 shows a mass m traveling with a velocity v . The basic variables used to describe the dynamic behavior of a translational mechanical system are the acceleration vector $\mathbf{a}$, the velocity vector $\mathbf{v}$, and the position vector $\mathbf{r}$. They are related by the time derivatives

$$\mathbf{a} = \frac{d\mathbf{v}}{dt} = \frac{d^2\mathbf{r}}{dt^2} \quad (5.1)$$

**FIGURE 5.1**

A mass traveling with a velocity v .

which can also be represented in the simple dot notation

$$\mathbf{a} = \dot{\mathbf{v}} = \ddot{\mathbf{r}} \quad (5.2)$$

Assume that the motion of the mass in [Figure 5.1](#) is under the influence of an externally applied force, and is constrained in only one direction. According to Newtonian mechanics, the resulting force f acting on the mass is equal to the time rate of change of momentum. For a constant mass, Newton's second law is expressed as

$$f = \frac{d}{dt}(mv) = m \frac{dv}{dt} = ma \quad (5.3)$$

Note that the acceleration a is absolute and must be measured with respect to an inertial reference frame. For ordinary systems at or near the surface of the Earth, the ground can be approximated as a reference for motion.

Mass elements store mechanical energy. The energy stored in a mass is kinetic energy if the mass is in motion. The kinetic energy is expressed as

$$T = \frac{1}{2}mv^2 \quad (5.4)$$

which implies that the mass stores kinetic energy as its velocity increases and releases kinetic energy as its velocity decreases. If a mass has a vertical displacement relative to a reference position, the energy stored in the mass is potential energy given by

$$V_g = mgh \quad (5.5)$$

where:

g is the gravitational acceleration (9.81 m/s² or 32.2 ft/s²)

h is the height measured from the reference position or datum to the center of mass

Subscript g is used to denote that the potential energy is associated with gravity

For rotational mechanical systems, the basic variables used to describe system dynamics are the angular acceleration vector $\boldsymbol{\alpha}$, the angular velocity vector $\boldsymbol{\omega}$, and the angular position vector $\boldsymbol{\theta}$. The direction of an angular vector can be determined using the right-hand rule, as shown in [Figure 5.2](#). The sense of rotation follows the curve of the four fingers, and the rotational vector points in the direction of the thumb. In this chapter, we consider the rigid bodies that are constrained to rotate about only one axis. Then, in scalar form, we have

$$\alpha = \frac{d\omega}{dt} = \frac{d^2\theta}{dt^2} \quad (5.6)$$

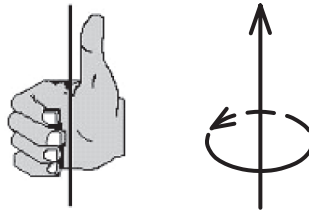


FIGURE 5.2
Right-hand rule.

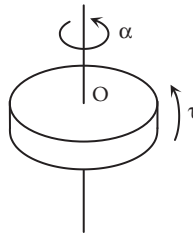


FIGURE 5.3
A disk rotating about an axis through a fixed point O.

or

$$\alpha = \dot{\omega} = \ddot{\theta} \quad (5.7)$$

Figure 5.3 shows a disk rotating about an axis through a fixed point O. The relation between the torque τ about the fixed point O and the angular acceleration α of the disk about O is

$$\tau = I_O \alpha \quad (5.8)$$

where I_O is the mass moment of inertia of the body about the fixed point O, and common units used are $\text{kg}\cdot\text{m}^2$ or $\text{slug}\cdot\text{ft}^2$. Similar to a translational mass, a rotational mass can store kinetic energy and potential energy. The kinetic energy for a rotational mass about a fixed point O is expressed as

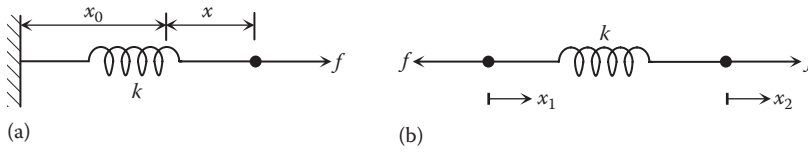
$$T = \frac{1}{2} I_O \omega^2 \quad (5.9)$$

The potential energy for a rotational mass has the same form as Equation 5.5.

5.1.2 Spring Elements

Figure 5.4a shows a translational spring element, which is fixed at one end and is subjected to a tensile (or compressive) force f at the other end. The spring has a free length x_0 , and the deflection of the spring caused by the force f is denoted by x . Assume that the spring is massless or of negligible mass. For a linear spring, Hooke's law states that

$$f = kx \quad (5.10)$$

**FIGURE 5.4**

A translational spring element with (a) one fixed end and (b) two free ends.

where k is the spring stiffness in units of N/m or lb/ft. If the spring is connected to a mass, due to Newton's third law, the force exerted on the mass by the spring has the same magnitude as f , but opposite in direction. When the two ends of a spring are displaced by x_1 and x_2 , as shown in Figure 5.4b, the forces at the two ends are equal in magnitude but opposite in direction. If $x_2 > x_1 > 0$, the spring is under elongation and the force applied to the spring is

$$f = kx_{\text{rel}} = k(x_2 - x_1) \quad (5.11)$$

where $x_{\text{rel}} = x_2 - x_1$ is the relative displacement between the two ends of the spring.

When a spring is stretched or compressed, potential energy is stored in the spring and is given by

$$V_e = \frac{1}{2}kx^2 \quad (5.12)$$

where subscript e denotes that the potential energy is associated with elastic elements.

For a torsional spring, as shown in Figure 5.5a, we have

$$\tau = K\theta \quad (5.13)$$

where:

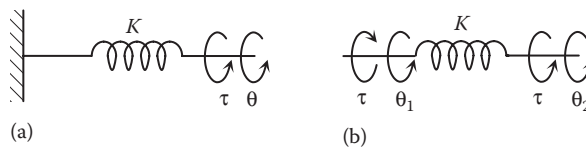
τ is the applied torque

K is the torsional spring stiffness in units of N·m/rad or ft·lb/rad

θ is the angular deformation of the spring

Figure 5.5b shows a torsional spring, with both ends twisted. Assume that θ_1 and θ_2 are the angular displacements of respective ends corresponding to the applied torque. If $\theta_2 > \theta_1 > 0$, then

$$\tau = K\theta_{\text{rel}} = K(\theta_2 - \theta_1) \quad (5.14)$$

**FIGURE 5.5**

A torsional spring element with (a) one fixed end and (b) two free ends.

and the spring is twisted in the counterclockwise direction when viewed from the right-hand side. The potential energy stored in a torsional spring element is expressed as

$$V_e = \frac{1}{2} K \theta^2 \quad (5.15)$$

5.1.3 Damper Elements

A spring element exerts a reaction force that is dependent on the relative displacement between two ends of the spring. In contrast, a force that depends on the relative velocity between two bodies is modeled by a damper element. Figure 5.6 shows a mass sliding on a fixed horizontal surface, where the two surfaces are separated by a film of liquid. The mass is subjected to a friction force generated between the two surfaces, and the friction caused by the liquid is called viscous damping. The direction of the damping force is opposite to the direction of the motion, and its magnitude depends on the nature of fluid flow between the two surfaces. The exact viscous damping force is complex; thus, for modeling in system dynamics, we use a linear relationship

$$f = bv \quad (5.16)$$

where the symbol b is used to denote the viscous damping coefficient in units of N·s/m or lb·s/ft. The damping force exerted on the mass in Figure 5.6 is to the left. Note that the symbol c is also often used to denote the viscous damping coefficient. Therefore, both b and c will be used interchangeably in this book.

The viscous friction can be modeled using a viscous damper (or a dashpot). The symbol in Figure 5.7a is the representation of a viscous damper, which is like a piston moving through a liquid-filled cylinder, as shown in Figure 5.7b. There are small holes in the piston through which the liquid flows as the parts move relative to each other. If $v_2 > v_1 > 0$, the right end of the damper moves to the right with respect to the left end. The force

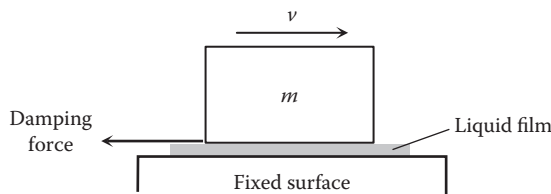


FIGURE 5.6

A mass sliding on a lubricated fixed surface.

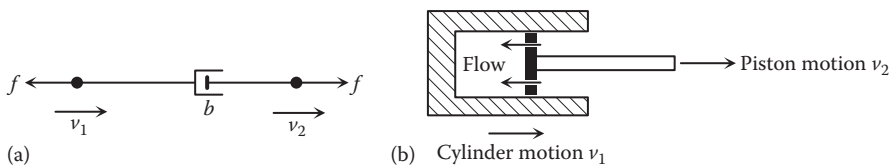
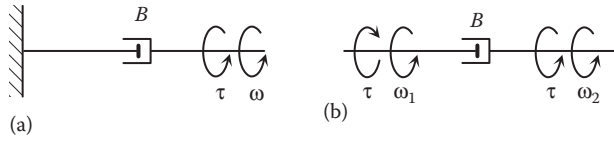


FIGURE 5.7

A viscous damper: (a) symbol and (b) physical system.

**FIGURE 5.8**

A rotational viscous damper with (a) one fixed end and (b) two free ends.

applied to the right end is dependent on the relative velocity $v_{\text{rel}} = v_2 - v_1$. The force has a magnitude of

$$f = bv_{\text{rel}} = b(v_2 - v_1) \quad (5.17)$$

and points to the right. Assume that the damper is massless or of negligible mass. Then, the forces at the two ends of the damper are equal in magnitude but opposite in direction.

For a torsional damper, as shown in Figure 5.8a, the linear relationship between the externally applied torque and the angular velocity is given by

$$\tau = B\omega \quad (5.18)$$

where B is the rotational viscous damping coefficient in units of N·m·s/rad or ft·lb·s/rad. The symbol in Figure 5.8b represents a rotational viscous damper, which can be used to model the viscous friction between two rotating surfaces separated by a film of liquid. If $\omega_2 > \omega_1 > 0$, the magnitude of the applied torque is

$$\tau = B\omega_{\text{rel}} = B(\omega_2 - \omega_1) \quad (5.19)$$

and the direction is as shown.

Note that the damping dissipates the energy of the system. Besides viscous damping, there are two other types of damping in engineering mechanics: Coulomb damping associated with dry friction and structural damping. The former will be discussed in Chapter 9, and the latter is beyond the scope of this text.

5.1.4 Equivalence

In many mechanical systems, multiple springs or dampers are used. In such cases, an equivalent spring stiffness constant or damping coefficient can be obtained to represent the combined elements.

Example 5.1: Springs in Parallel

Consider a system of two springs, k_1 and k_2 , in parallel, as shown in Figure 5.9. Prove that the system is equivalent to a single spring whose stiffness is

$$k_{\text{eq}} = k_1 + k_2$$

Proof

Because of parallel interconnection, the bottom ends of the springs are attached to the same fixed body, and their top ends are also attached to a common body. This implies that both springs have the same deflection x . Assume that the forces applied

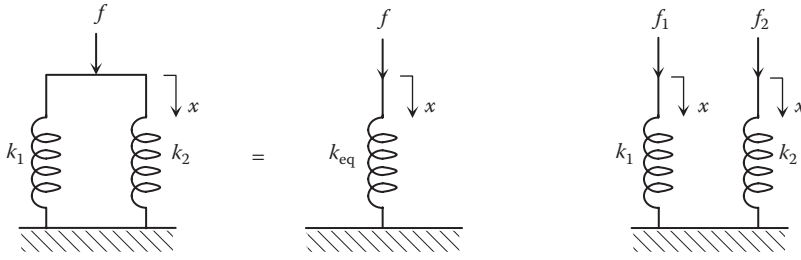


FIGURE 5.9
Equivalence for two springs in parallel.

to the two springs are f_1 and f_2 , respectively. Because the system is in static equilibrium, the total force is given by

$$f = f_1 + f_2 = k_1x + k_2x = (k_1 + k_2)x$$

Comparing it with the equivalent system,

$$f = k_{\text{eq}}x$$

we obtain the equivalent spring stiffness

$$k_{\text{eq}} = k_1 + k_2$$

The result can be extended to n springs. For a system of n springs in parallel, the equivalent spring stiffness k_{eq} is equal to the sum of all the individual spring stiffness coefficients k_i :

$$k_{\text{eq}} = k_1 + k_2 + \cdots + k_n$$

Example 5.2: Springs in Series

Consider a system of two springs, k_1 and k_2 , in series, as shown in [Figure 5.10](#). Prove that the equivalent spring stiffness of the system is

$$k_{\text{eq}} = \frac{k_1 k_2}{k_1 + k_2}$$

Proof

Because both springs are in static equilibrium, they are subjected to the same force f . Assume that the two springs are deformed by x_1 and x_2 , respectively. The total deformation of the system is given by

$$x = x_1 + x_2 = \frac{f}{k_1} + \frac{f}{k_2} = f \left(\frac{1}{k_1} + \frac{1}{k_2} \right)$$

For the equivalent system, the deformation is

$$x = \frac{f}{k_{\text{eq}}}$$

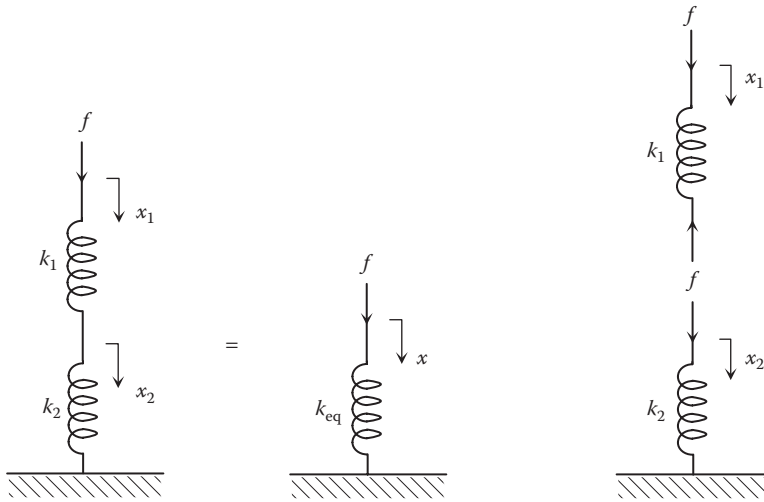


FIGURE 5.10
Equivalence for two springs in series.

Thus,

$$\frac{1}{k_{\text{eq}}} = \frac{1}{k_1} + \frac{1}{k_2}$$

or

$$k_{\text{eq}} = \frac{k_1 k_2}{k_1 + k_2}$$

The result can also be extended to n springs. For a system of n springs in series, the reciprocal of the equivalent spring stiffness k_{eq} is equal to the sum of all the reciprocals of the individual spring stiffness coefficients k_i :

$$\frac{1}{k_{\text{eq}}} = \frac{1}{k_1} + \frac{1}{k_2} + \cdots + \frac{1}{k_n}$$

The above-mentioned two examples show how one can derive the equivalent spring stiffness for springs connected in parallel or in series. For a system of dampers, the equivalent damping coefficient can be derived using the same logic and similar steps.

Springs are the most familiar elastic elements. However, many engineering applications involving elastic elements do not contain springs but other mechanical elements, such as beams and rods, which can be modeled as springs. The equivalent spring constants can be determined using the results from the study of mechanics of materials [2, 12].

Example 5.3: Equivalent Spring Constant of a Cantilever Beam

Consider a uniform cantilever beam of length L , width b , and thickness h in [Figure 5.11](#). Assume that a force f is applied to the free end of the beam, and the corresponding deflection is x . Derive the equivalent spring constant k_{eq} .

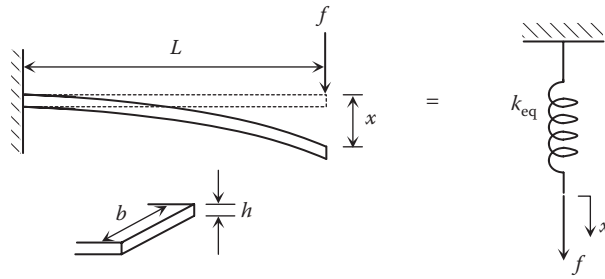


FIGURE 5.11
A beam in bending under a transverse force.

Solution

The force–deflection relation of a cantilever beam can be found in mechanics of materials references. The relation is

$$x = \frac{L^3}{3EI_A} f$$

where:

x is the deflection at the free end of the beam

f is the force applied at the free end of the beam

E is the modulus of elasticity of beam material

I_A is the area moment of inertia about the beam's longitudinal axis

For a beam having a rectangular cross-section, with width b and thickness h , the area moment of inertia is

$$I_A = \frac{bh^3}{12}$$

Thus, the force–deflection relation reduces to

$$x = \frac{4L^3}{Eb h^3} f$$

For the equivalent system, the force–deflection relation is

$$x = \frac{f}{k_{eq}}$$

Thus, the equivalent spring stiffness is

$$k_{eq} = \frac{Eb h^3}{4L^3}$$

PROBLEM SET 5.1

1. If the 50-kg block in [Figure 5.12](#) is released from rest at *A*, determine its kinetic energy and velocity after it slides 5 m down the plane. Assume that the plane is smooth.
2. Repeat Problem 1 if the coefficient of kinetic friction between the block and the plane is $\mu_k = 0.15$.
3. The ball in [Figure 5.13](#) has a mass of 6 kg and is fixed to a rod having a negligible mass. Assume that the ball is released from rest when $\theta = 60^\circ$.
 - a. Determine the gravitational potential energy of the ball when $\theta = 60^\circ$ and 30° . The datum is shown in [Figure 5.13](#).
 - b. Determine the kinetic energy and the velocity of the ball when $\theta = 30^\circ$.
4. The 6-kg slender rod in [Figure 5.14](#) is released from rest when $\theta = 60^\circ$.
 - a. Determine the gravitational potential energy of the rod when $\theta = 60^\circ$ and 30° . The datum is shown in [Figure 5.14](#).
 - b. Determine the kinetic energy and the angular velocity of the rod when $\theta = 30^\circ$. The mass moment of inertia of the slender rod about the fixed point *O* is $I_O = mL^2/3$, where *L* is the length of the rod.
5. Determine the elastic potential energy of the system shown in [Figure 5.15](#) if the 10-kg block moves downward a distance of 0.05 m. Assume that the block is originally in static equilibrium.
6. If the disk in [Figure 5.16](#) rotates in the clockwise direction by 10° , determine the elastic potential energy of the system. Assume that the springs are originally undeformed.
7. Determine the equivalent spring constant for the system shown in [Figure 5.17](#).

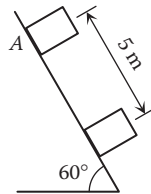


FIGURE 5.12
Problem 1.

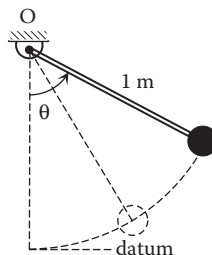


FIGURE 5.13
Problem 3.

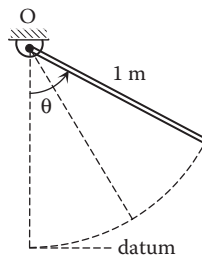


FIGURE 5.14
Problem 4.

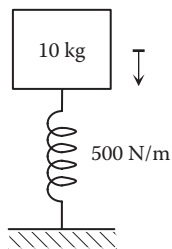


FIGURE 5.15
Problem 5.

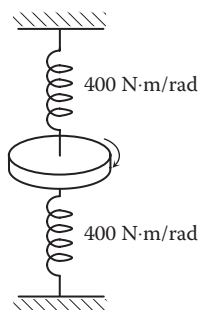


FIGURE 5.16
Problem 6.

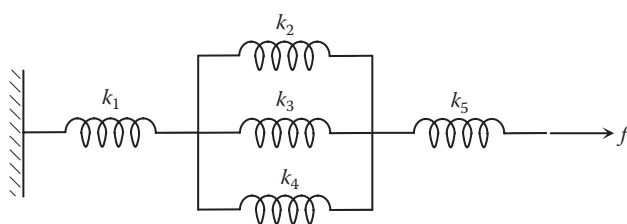


FIGURE 5.17
Problem 7.

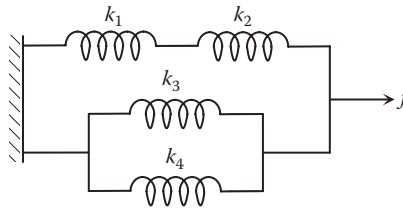


FIGURE 5.18
Problem 8.

8. Determine the equivalent spring constant for the system shown in [Figure 5.18](#).
9. Derive the spring constant expression for the axially loaded bar shown in [Figure 5.19](#). Assume that the cross-sectional area is A and the modulus of elasticity of the material is E .
10. The uniform circular shaft in [Figure 5.20](#) acts as a torsional spring. Assume that the elastic shear modulus of the material is G . Derive the equivalent spring constant corresponding to a pair of torques applied at the two free ends.
11. A rod is made of two uniform sections, as shown in [Figure 5.21](#). The two sections are made of the same material, and the modulus of elasticity of the rod material is E . The areas for the two sections are A_1 and A_2 , respectively. Derive the equivalent spring constant corresponding to a tensile force f applied at the free end.
12. Derive the spring constant expression for the fixed-fixed beam in [Figure 5.22](#). The Young's modulus of the material is E and the moment of inertia of cross-sectional area is I . Assume that the force f and the deflection x are at the center of the beam.

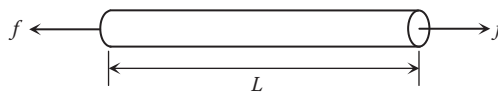


FIGURE 5.19
Problem 9.

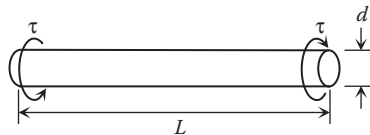


FIGURE 5.20
Problem 10.

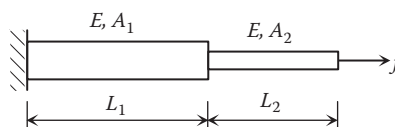


FIGURE 5.21
Problem 11.

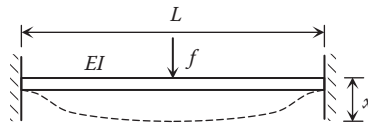


FIGURE 5.22
Problem 12.

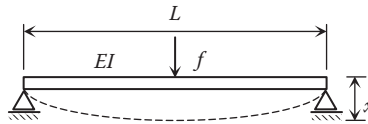


FIGURE 5.23
Problem 13.

13. Derive the spring constant expression for the simply supported beam in [Figure 5.23](#). The Young's modulus of the material is E and the moment of inertia of cross-sectional area is I . Assume that the force f and the deflection x are at the center of the beam.
14. Derive the spring constant expression for the simply supported beam in [Figure 5.24](#). The Young's modulus of the material is E and the moment of inertia of cross-sectional area is I . Assume that the applied load f is anywhere between the supports.
15. Determine the equivalent damping coefficient for the system shown in [Figure 5.25](#).
16. Determine the equivalent damping coefficient for the system shown in [Figure 5.26](#).

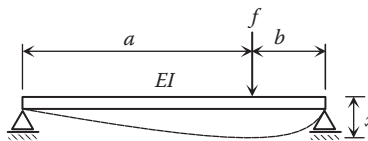


FIGURE 5.24
Problem 14.

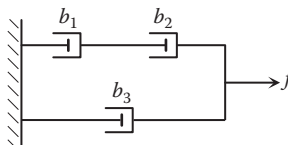


FIGURE 5.25
Problem 15.

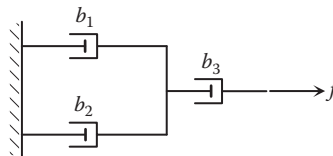


FIGURE 5.26
Problem 16.

5.2 Translational Systems

With appropriate simplifying assumptions, a translational mechanical system can be modeled as a system of interconnected mechanical elements. The dynamic behavior of the system must obey the physical laws, and the dynamic equations of motion can be obtained by applying these physical laws, such as Newton's second law or D'Alembert's principle. The number of equations of motion is determined by the number of degrees of freedom of the system.

5.2.1 Degrees of Freedom

The number of degrees of freedom of a dynamic system is defined as the number of independent generalized coordinates that specify the configuration of the system. Generalized coordinates need not be restricted only to the actual position coordinates, which are physical coordinates. They could be anything, for example, position coordinates, translational displacement, rotational displacement, pressure, voltage, and current. The generalized coordinates of a system need not be of the same type.

Figure 5.27a shows a translational mechanical system, in which the mass m moves in the horizontal direction and x is the displacement measured from the static equilibrium position of the mass. The displacement x is the generalized coordinate, and the number of degrees of freedom of the system is 1. When a pendulum consisting of a massless rod of length L and a point mass of M is attached to the block of mass m (Figure 5.27b), one displacement coordinate x is not enough to describe the motion of the system. On the one hand, the pendulum moves together with the block in the horizontal direction. On the other hand, the pendulum rotates and the rotational motion can be described using an angular displacement θ . Thus, for the system in Figure 5.27b, x and θ are the two independent generalized coordinates and the number of degrees of freedom of the system is 2.

When generalized coordinates are independent, they are equal in number to the degrees of freedom. If generalized coordinates are dependent, the number of degrees of freedom is the difference between the number of dependent coordinates and the number of constraints. For instance, to describe the motion of the pendulum in Figure 5.28, we can use the rectangular displacements x and y , instead of the angular displacement θ . However,

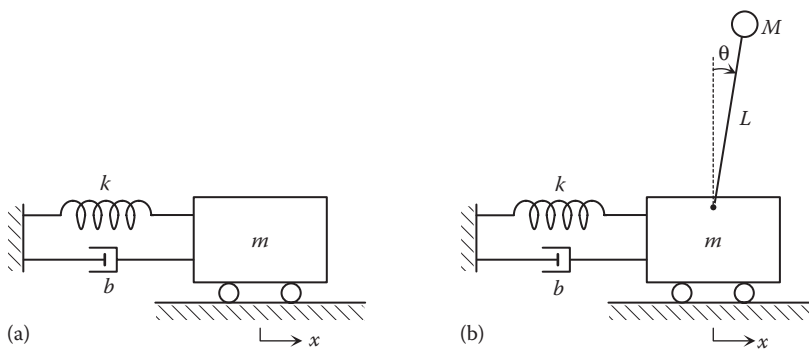


FIGURE 5.27

A mechanical system with (a) displacement as the generalized coordinate and (b) mixed types of generalized coordinates.

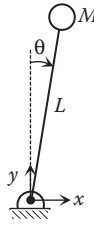


FIGURE 5.28
Independence of generalized coordinates.

those two rectangular displacements are related by the constraint $x^2 + y^2 = L^2$. Thus, the number of degrees of freedom is 1 (i.e., number of dependent generalized coordinates—number of constraints), which is the same as the result obtained by using the angular displacement as the generalized coordinate.

5.2.2 Newton's Second Law

Newton's second law states that the acceleration of a mass is proportional to the resultant force vector acting on it and is in the direction of this force. Assume that the translational motion of a particle or a rigid body is restricted in a plane. For a particle, which is a mass of negligible dimensions, Newton's second law can be expressed in vector form

$$\sum \mathbf{F} = m\mathbf{a} \quad (5.20)$$

or in scalar form

$$\sum F_x = ma_x, \quad \sum F_y = ma_y \quad (5.21)$$

where $\sum F_x$ and $\sum F_y$ are summations of the applied forces decomposed along the x and y directions, respectively, and a_x and a_y are the x and y components of the acceleration of the particle, respectively.

For a rigid body, Newton's second law is given by

$$\sum \mathbf{F} = m\mathbf{a}_C \quad (5.22)$$

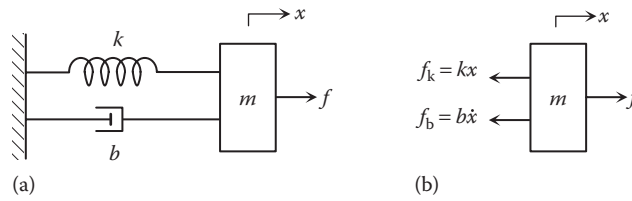
or

$$\sum F_x = ma_{Cx}, \quad \sum F_y = ma_{Cy} \quad (5.23)$$

where subscript C denotes the center of mass. In many engineering applications, the gravity field is considered to be uniform and the center of mass coincides with the center of gravity.

5.2.3 Free-Body Diagrams

To apply Newton's second law to a mechanical system, it is useful to draw a free-body diagram for each mass in the system, showing all external forces. The non-input forces can be described in terms of displacements or velocities by using the expressions associated with the basic spring or damper elements. Drawing correct free-body diagrams is the most important step in analyzing mechanical systems by the force/moment approach (as opposed to the energy approach).

**FIGURE 5.29**

A mass–spring–damper system: (a) physical system and (b) free-body diagram.

Let us consider a simple system consisting of a block of mass m , a spring of stiffness k , and a viscous damper of viscous damping coefficient b . Figure 5.29 shows the physical mass–spring–damper system and the free-body diagram drawn for the mass. The motion of the system can be described using the displacement variable x , which is chosen as the generalized coordinate. The positive direction is the direction shown by the arrow next to the displacement x . Assume that the positive direction is to the right, as shown. This sign convention implies that the displacement x , velocity $\dot{x}$, and acceleration $\ddot{x}$ are positive to the right.

Three forces included in the free-body diagram are the applied force f , the force exerted by the spring f_k , and the force exerted by the damper f_b . The magnitudes of the forces are shown in the free-body diagram, and their physical directions are indicated by the arrows. The force f is externally applied to the mass–spring–damper system, and the positive direction is given to the right. To determine the forces f_k and f_b , we can imagine the mass to be displaced along the positive direction, $x > 0$. Thus, the spring is in tension, and there is a spring force $f_k = kx$ applied to the mass. The force exerted by the spring is to the left, as it tends to restore to the undeformed position. Similarly, the assumption of $\dot{x} > 0$ indicates that the mass moves to the right with a velocity $\dot{x}$. Remember that the damping force for a moving mass is always opposite to the direction of motion. Thus, the force exerted by the damper is to the left, and the magnitude is $f_b = b\dot{x}$.

The previous analysis shows that an assumption about the motion of all masses in a mechanical system must be made to draw the free-body diagrams. It is customary to assume that all displacements are in the assumed positive directions when determining the proper magnitudes and directions for the forces. Applying Newton’s second law to the correct free-body diagrams leads to differential equations of motion, which can be converted to other system representations, such as the transfer function form and the state-space form.

Example 5.4: A Single-Degree-of-Freedom Mass–Spring–Damper System

Consider the simple mass–spring–damper system subjected to an input force f , as shown in Figure 5.29a.

- Apply Newton’s second law to derive the differential equation of motion.
- Determine the transfer function. Assume that the system output is the displacement x and the initial conditions are $x(0) = 0$ and $\dot{x}(0) = 0$.
- Determine the state-space representation. Assume that the system output is the displacement x and the state variables are $x_1 = x$ and $x_2 = \dot{x}$.
- Use Simulink and Simscape to construct block diagrams to find the displacement output $x(t)$ of the system subjected to an applied force $f(t) = 10u(t)$, where $u(t)$ is the unit-step function. The parameter values are $m = 1$ kg, $b = 2$ N·s/m, and $k = 5$ N/m. Assume zero initial conditions.

Solution

- a. Let us choose the displacement of the mass as the coordinate x . The free-body diagram of the mass is shown in [Figure 5.29b](#). Applying Newton's second law in the x direction gives

$$+\rightarrow x: \sum F_x = ma_x$$

$$f - kx - b\dot{x} = m\ddot{x}$$

which can be rearranged into the standard input–output differential equation form

$$m\ddot{x} + b\dot{x} + kx = f$$

- b. Taking the Laplace transform of both sides of the preceding equation with zero initial conditions results in

$$(ms^2 + bs + k)X(s) = F(s)$$

Thus, the transfer function relating the input $f(t)$ to the output $x(t)$ is

$$\frac{X(s)}{F(s)} = \frac{1}{ms^2 + bs + k}$$

- c. As specified, the state, the input, and the output are

$$\mathbf{x} = \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix} = \begin{Bmatrix} x \\ \dot{x} \end{Bmatrix}, \quad u = f, \quad y = x$$

The state-variable equations are thus formed as

$$\dot{x}_1 = \dot{x} = x_2$$

$$\dot{x}_2 = \ddot{x} = -\frac{k}{m}x - \frac{b}{m}\dot{x} + \frac{1}{m}f = -\frac{k}{m}x_1 - \frac{b}{m}x_2 + \frac{1}{m}u$$

The output equation is

$$y = x = x_1$$

Writing the state-variable equations and the output equation in matrix form yields

$$\begin{Bmatrix} \dot{x}_1 \\ \dot{x}_2 \end{Bmatrix} = \begin{bmatrix} 0 & 1 \\ -\frac{k}{m} & -\frac{b}{m} \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix} + \begin{bmatrix} 0 \\ \frac{1}{m} \end{bmatrix} u$$

$$y = \begin{bmatrix} 1 & 0 \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix} + 0 \cdot u$$

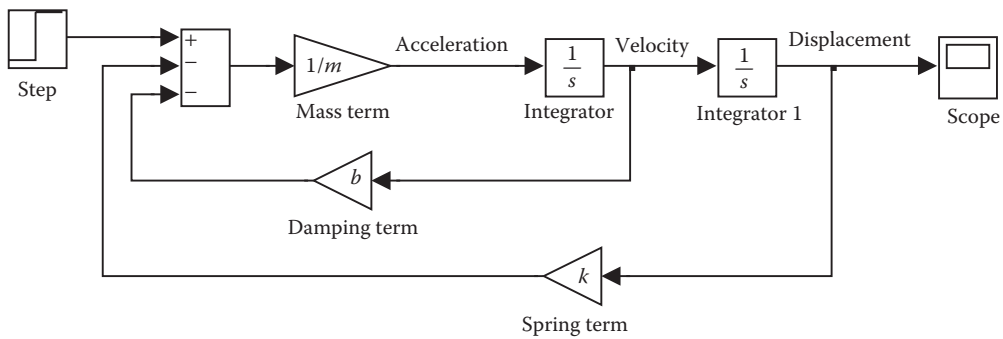


FIGURE 5.30
Simulink block diagram corresponding to Example 5.4.

- d. We start with building a Simulink block diagram by using the mathematical model obtained in Part (a). Solving for the highest derivative of the output x gives

$$\ddot{x} = \frac{1}{m}(f - kx - b\dot{x})$$

which can be represented using the block diagram shown in Figure 5.30. Two Integrator blocks are used to form the velocity $\dot{x}$ and the displacement x , both of which are fed back to form the acceleration $\ddot{x}$. Note that a step input causes the motion of the system. Double-click the block with the name Step, and type 0 for the Step time and 10 for the Final value to define the input $f(t) = 10u(t)$.

We can also build a Simscape block diagram to simulate the physical system shown in Figure 5.29a. The mass, translational damper, and translational spring blocks can be found in Simscape/Foundation Library/Mechanical/Translational Elements. In the same library, the Mechanical Translational Reference block is used to represent a rigidly clamped end. To apply a force input, the Ideal Force Source block is included, which can be found in .../Mechanical/Mechanical Sources. To obtain a displacement output, the Ideal Translational Motion Sensor block is included, which can be found in .../Mechanical/Mechanical Sensors. Figure 5.31 is the resulting Simscape block diagram. Note that port symbols (e.g., C for “case”, R for “rod”, P for “position”, V for “velocity”, and S for “signal”) are useful for making a correct connection.

Define the values of the parameters m , b , and k in the MATLAB Command window. Run both simulations, and the same curve, as shown in Figure 5.32, can be obtained, which is the resulting displacement output $x(t)$ of the mass–spring–damper system in Figure 5.29a, subjected to a step input force. More examples of using Simulink and Simscape to build a mechanical system will be given in Section 5.6.

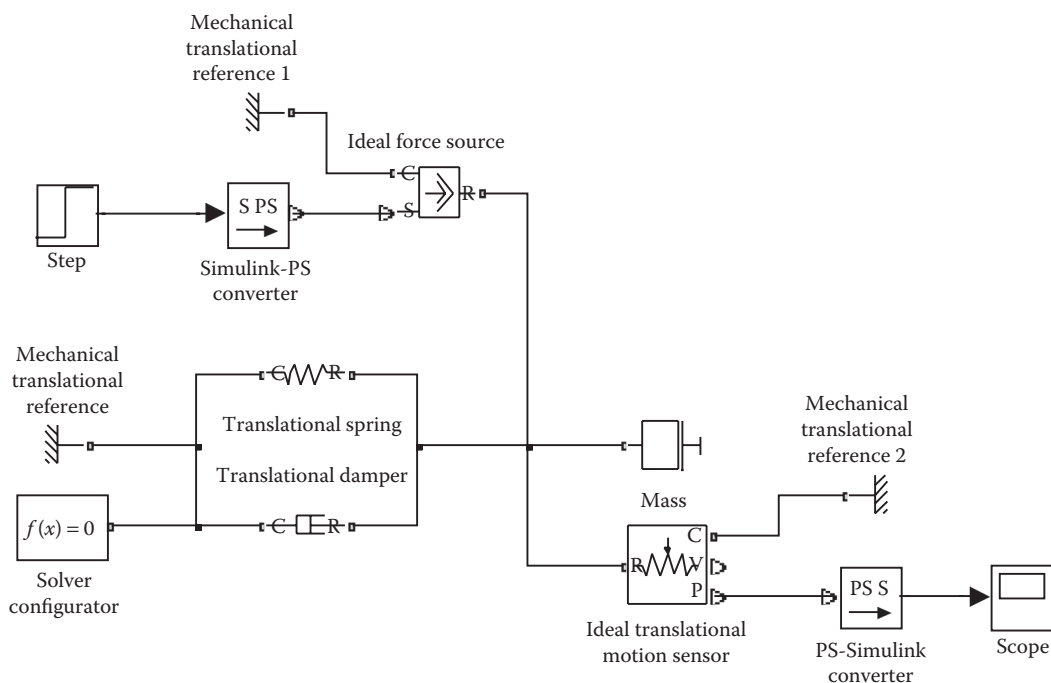


FIGURE 5.31
Simscape block diagram corresponding to Example 5.4.

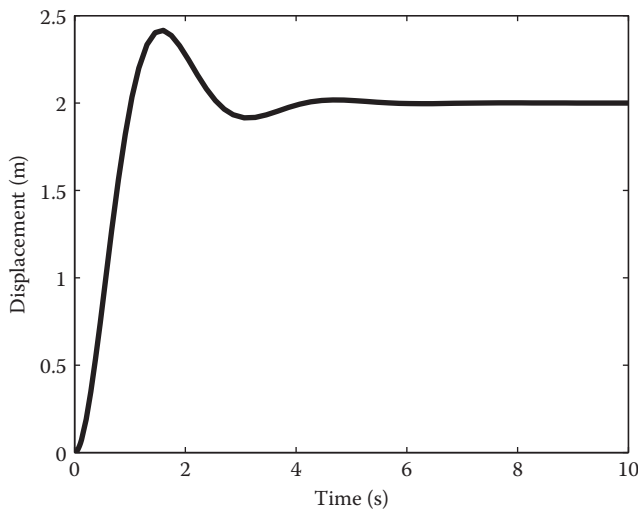


FIGURE 5.32
Displacement output $x(t)$ of the mechanical system in Example 5.4.

5.2.4 Static Equilibrium Position and Coordinate Reference

In Example 5.4, we specified the static equilibrium position as the coordinate origin. For this mass–spring–damper system moving only in the horizontal direction, the mass is in equilibrium when the spring is at its free length. Note that it is advantageous to choose the static equilibrium position as the coordinate origin, because this choice can simplify the equation of motion by eliminating static forces. The advantage is obvious when the motion along the vertical direction is involved. The following example shows that the gravity term does not enter the governing differential equation if the displacement is measured from the static equilibrium.

Consider the mass–spring system shown in [Figure 5.33a](#), where the mass is assumed to move only in the vertical direction. The free length of the spring is y_0 . Due to gravity, the spring is stretched by δ_{st} when the mass is in static equilibrium and $mg = k\delta_{st}$ ([Figure 5.33b](#)). Imagine the mass to be displaced downward by a distance of x . If we choose the undeformed position in [Figure 5.33a](#) as the origin of the coordinate y , applying Newton’s second law to the free-body diagram in [Figure 5.33c](#) gives

$$\begin{aligned} +\downarrow y: \sum F_y &= ma_y \\ mg - ky &= m\ddot{y} \\ m\ddot{y} + ky &= mg \end{aligned}$$

Note that the gravity term mg appears in the equation of motion. Now, let us choose the static equilibrium position in [Figure 5.33b](#) as the origin of the coordinate x . The equation of motion is

$$\begin{aligned} +\downarrow x: \sum F_x &= ma_x \\ mg - k(x + \delta_{st}) &= m\ddot{x} \\ m\ddot{x} + kx &= 0 \end{aligned}$$

where the gravity term mg and the static spring force $k\delta_{st}$ cancel out, resulting in a simpler equation.

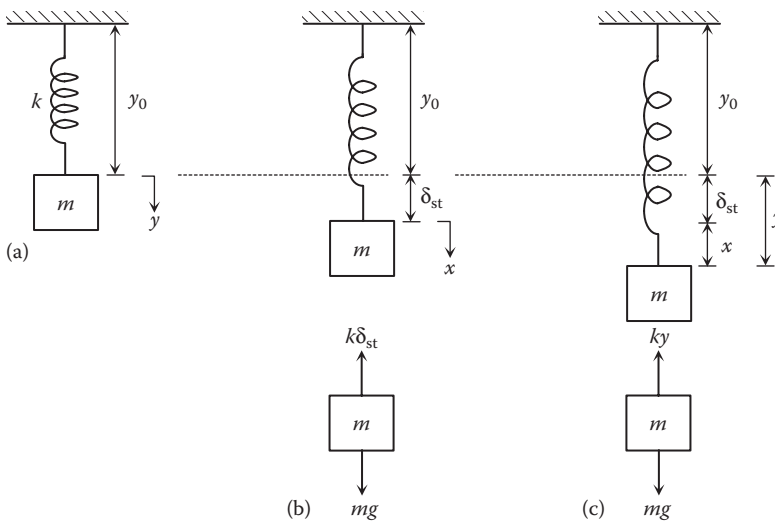


FIGURE 5.33

Choice of coordinate origins for a mass–spring system: (a) undeformed position, (b) static equilibrium position, and (c) dynamic position.

Example 5.5: A Two-Degree-of-Freedom Quarter-Car Model

Consider a quarter-car model shown in Figure 5.34a, where m_1 is the mass of one-fourth of the car body and m_2 is the mass of the wheel–tire–axle assembly. The spring k_1 represents the elasticity of the suspension and the spring k_2 represents the elasticity of the tire. $z(t)$ is the displacement input due to the surface of the road. The actuator force, f , applied between the car body and the wheel–tire–axle assembly, is controlled by feedback and represents the active components of the suspension system.

- Draw the necessary free-body diagrams, and derive the differential equations of motion.
- Determine the state-space representation. Assume that the displacements of the two masses, x_1 and x_2 , are the outputs and the state variables are $x_1 = x_1$, $x_2 = x_2$, $x_3 = \dot{x}_1$, and $x_4 = \dot{x}_2$.
- ▲ The parameter values are $m_1 = 290$ kg, $m_2 = 59$ kg, $b_1 = 1000$ N·s/m, $k_1 = 16,182$ N/m, and $k_2 = 19,000$ N/m. Use MATLAB commands to define the system in the state-space form and then convert it to the transfer function form. Assume that all the initial conditions are zero.

Solution

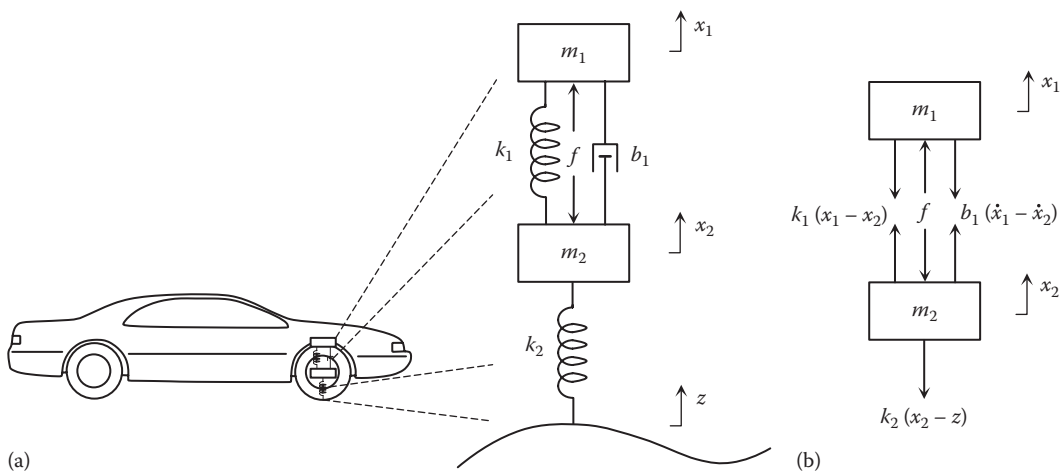
- We choose the displacements of the two masses x_1 and x_2 as the generalized coordinates. The static equilibrium positions of m_1 and m_2 are set as the coordinate origins. Assume

$$x_1 > x_2 > z > 0$$

which implies that the springs are in tension and

$$\dot{x}_1 > \dot{x}_2 > \dot{z} > 0$$

The free-body diagrams of m_1 and m_2 are shown in Figure 5.34b. According to the assumption, the mass m_1 moves faster than the mass m_2 , and the elongation of the spring k_1 is $x_1 - x_2$. The force exerted by the spring k_1 on the mass m_1 is downward as it tends to restore to the undeformed position. Because of

**FIGURE 5.34**

A quarter-car model: (a) physical system and (b) free-body diagram.

Newton's third law, the force exerted by the spring k_1 on the mass m_2 has the same magnitude, but opposite in direction. Other spring forces and damping forces can be determined using the same logic. Note that the gravitational forces, m_1g and m_2g , are not included in the free-body diagrams.

Applying Newton's second law to the masses m_1 and m_2 , respectively, gives

$$\begin{aligned} +\uparrow x: \sum F_x &= ma_x \\ f - k_1(x_1 - x_2) - b_1(\dot{x}_1 - \dot{x}_2) &= m_1\ddot{x}_1 \\ -f + k_1(x_1 - x_2) + b_1(\dot{x}_1 - \dot{x}_2) - k_2(x_2 - z) &= m_2\ddot{x}_2 \end{aligned}$$

Rearranging the equations into the standard input–output form,

$$\begin{aligned} m_1\ddot{x}_1 + b_1\dot{x}_1 - b_1\dot{x}_2 + k_1x_1 - k_1x_2 &= f \\ m_2\ddot{x}_2 - b_1\dot{x}_1 + b_1\dot{x}_2 - k_1x_1 + (k_1 + k_2)x_2 &= -f + k_2z \end{aligned}$$

which can be expressed in second-order matrix form (Section 4.1) as

$$\begin{bmatrix} m_1 & 0 \\ 0 & m_2 \end{bmatrix} \begin{Bmatrix} \ddot{x}_1 \\ \ddot{x}_2 \end{Bmatrix} + \begin{bmatrix} b_1 & -b_1 \\ -b_1 & b_1 \end{bmatrix} \begin{Bmatrix} \dot{x}_1 \\ \dot{x}_2 \end{Bmatrix} + \begin{bmatrix} k_1 & -k_1 \\ -k_1 & k_1 + k_2 \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix} = \begin{bmatrix} 1 & 0 \\ -1 & k_2 \end{bmatrix} \begin{Bmatrix} f \\ z \end{Bmatrix}$$

- b. Note that the inputs to the system are the actuator force f and the road surface irregularity z . The state, the input, and the output vectors are

$$\mathbf{x} = \begin{Bmatrix} x_1 \\ x_2 \\ x_3 \\ x_4 \end{Bmatrix} = \begin{Bmatrix} x_1 \\ x_2 \\ \dot{x}_1 \\ \dot{x}_2 \end{Bmatrix}, \quad \mathbf{u} = \begin{Bmatrix} f \\ z \end{Bmatrix}, \quad \mathbf{y} = \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix}$$

The state-variable equations are then obtained as

$$\begin{aligned} \dot{x}_1 &= x_3 \\ \dot{x}_2 &= x_4 \\ \dot{x}_3 &= \ddot{x}_1 = -\frac{k_1}{m_1}x_1 + \frac{k_1}{m_1}x_2 - \frac{b_1}{m_1}\dot{x}_1 + \frac{b_1}{m_1}\dot{x}_2 + \frac{f}{m_1} \\ &= -\frac{k_1}{m_1}x_1 + \frac{k_1}{m_1}x_2 - \frac{b_1}{m_1}x_3 + \frac{b_1}{m_1}x_4 + \frac{1}{m_1}u_1 \\ \dot{x}_4 &= \ddot{x}_2 = \frac{k_1}{m_2}x_1 - \frac{k_1 + k_2}{m_2}x_2 + \frac{b_1}{m_2}\dot{x}_1 - \frac{b_1}{m_2}\dot{x}_2 - \frac{f}{m_2} + \frac{k_2}{m_2}z \\ &= \frac{k_1}{m_2}x_1 - \frac{k_1 + k_2}{m_2}x_2 + \frac{b_1}{m_2}x_3 - \frac{b_1}{m_2}x_4 - \frac{1}{m_2}u_1 + \frac{k_2}{m_2}u_2 \end{aligned}$$

The output equation is

$$\mathbf{y} = \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix}$$

Thus, the state-space representation is

$$\begin{Bmatrix} \dot{x}_1 \\ \dot{x}_2 \\ \dot{x}_3 \\ \dot{x}_4 \end{Bmatrix} = \begin{bmatrix} 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 1 \\ -\frac{k_1}{m_1} & \frac{k_1}{m_1} & -\frac{b_1}{m_1} & \frac{b_1}{m_1} \\ \frac{k_1}{m_2} & -\frac{k_1+k_2}{m_2} & \frac{b_1}{m_2} & -\frac{b_1}{m_2} \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \\ x_3 \\ x_4 \end{Bmatrix} + \begin{bmatrix} 0 & 0 \\ 0 & 0 \\ \frac{1}{m_1} & 0 \\ -\frac{1}{m_2} & \frac{k_2}{m_2} \end{bmatrix} \begin{Bmatrix} u_1 \\ u_2 \end{Bmatrix}$$

$$\mathbf{y} = \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \\ x_3 \\ x_4 \end{Bmatrix} + \begin{bmatrix} 0 & 0 \\ 0 & 0 \end{bmatrix} \begin{Bmatrix} u_1 \\ u_2 \end{Bmatrix}$$

c.  The following is the MATLAB session:

```
>> m1 = 290;
>> m2 = 59;
>> b1 = 1000;
>> k1 = 16182;
>> k2 = 19000;
>> A = [0 0 1 0;
        0 0 0 1;
        -k1/m1 k1/m1 -b1/m1 b1/m1;
        k1/m2 -(k1+k2)/m2 b1/m2 -b1/m2];
>> B = [0 0; 0 0; 1/m1 0; -1/m2 k2/m2];
>> C = [1 0 0 0; 0 1 0 0];
>> D = zeros(2,2);
>> sys_ss = ss(A,B,C,D);
>> sys_tf = tf(sys_ss);
```

The command `tf` returns two transfer functions from the input #1 (i.e., f) to the two outputs x_1 and x_2 :

$$\frac{X_1(s)}{F(s)} = \frac{0.003448s^2 + 1.11}{s^4 + 20.4s^3 + 652.1s^2 + 1110s + 17,970}$$

$$\frac{X_2(s)}{F(s)} = \frac{-0.01695s^2}{s^4 + 20.4s^3 + 652.1s^2 + 1110s + 17,970}$$

and another two transfer functions from the input #2 (i.e., z) to the outputs x_1 and x_2 :

$$\frac{X_1(s)}{Z(s)} = \frac{1110s + 17,970}{s^4 + 20.4s^3 + 652.1s^2 + 1110s + 17,970}$$

$$\frac{X_2(s)}{Z(s)} = \frac{322s^2 + 1110s + 17,970}{s^4 + 20.4s^3 + 652.1s^2 + 1110s + 17,970}$$

Note that the system has two inputs and two outputs. Thus, there are a total of four transfer functions, which can be formed as a 2×2 transfer matrix

$$\mathbf{G}(s) = \begin{bmatrix} G_{11}(s) & G_{12}(s) \\ G_{21}(s) & G_{22}(s) \end{bmatrix} = \begin{bmatrix} \frac{X_1(s)}{F(s)} & \frac{X_1(s)}{Z(s)} \\ \frac{X_2(s)}{F(s)} & \frac{X_2(s)}{Z(s)} \end{bmatrix}$$

Note that the gravity terms in Example 5.5 do not appear in the equations of motion, because the static equilibrium positions are chosen as the coordinate origins. Two independent coordinates, x_1 and x_2 , are required to specify the system dynamics. Such a system is called a two-degree-of-freedom system, which is a special case of multiple-degree-of-freedom systems.

5.2.5 Massless Junctions

A system of massless junctions is a system of springs and dampers without any masses. The differential equations of motion for such a system can be derived using Newton's second law and simply letting the masses be zero at the junctions.

Example 5.6: A Two-Degree-of-Freedom System with Massless Junctions

Consider the system of massless junctions shown in Figure 5.35a. An external force f is applied to the junction A. Draw the free-body diagrams and derive the differential equations of motion.

Solution

Two massless junctions, A and B, are included in this system. We choose the displacements of the two junctions as the generalized coordinates, which are denoted by x_1 and x_2 . Assume that $x_1 > x_2 > 0$. This implies that the two springs are in extension. The free-body diagrams at the two massless junctions are shown in Figure 5.35b. Applying Newton's second law to each massless junction gives

$$+ \rightarrow x: \sum F_x = ma_x = 0$$

$$\text{A: } -k_1(x_1 - x_2) - b_1(\dot{x}_1 - \dot{x}_2) + f = 0$$

$$\text{B: } k_1(x_1 - x_2) - k_2x_2 + b_1(\dot{x}_1 - \dot{x}_2) = 0$$

The equations can be rearranged as

$$b_1\dot{x}_1 - b_1\dot{x}_2 + k_1x_1 - k_1x_2 = f$$

$$-b_1\dot{x}_1 + b_1\dot{x}_2 - k_1x_1 + (k_1 + k_2)x_2 = 0$$

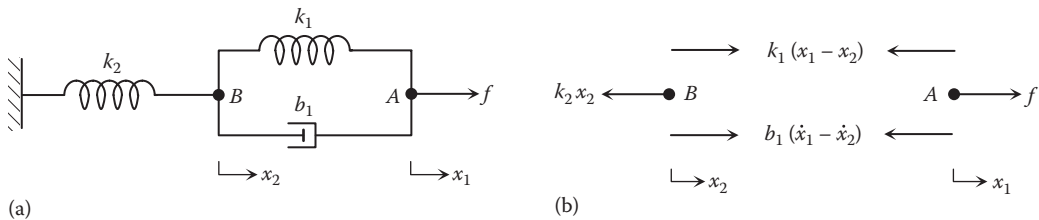
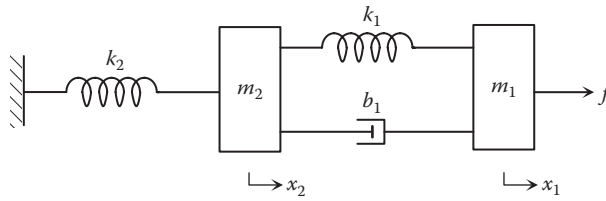


FIGURE 5.35

A two-degree-of-freedom system with massless junctions: (a) physical system and (b) free-body diagram.

**FIGURE 5.36**

System obtained by replacing the massless junctions in Figure 5.35 by masses.

or in matrix form

$$\begin{bmatrix} b_1 & -b_1 \\ -b_1 & b_1 \end{bmatrix} \begin{Bmatrix} \dot{x}_1 \\ \dot{x}_2 \end{Bmatrix} + \begin{bmatrix} k_1 & -k_1 \\ -k_1 & k_1 + k_2 \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix} = \begin{Bmatrix} f \\ 0 \end{Bmatrix}$$

Note that the system in Example 5.6 is a two-degree-of-freedom system. The dynamic behavior of the system is described by two first-order differential equations of motion, and thus, it is a second-order system. If the two massless junctions are replaced by two masses, as shown in Figure 5.36, the resulting system is still a two-degree-of-freedom system, but a fourth-order system. The reader can derive the differential equations of motion for the new system in Figure 5.36 as an exercise.

In Examples 5.5 and 5.6, the differential equations of motion are also given in second-order matrix form as $\mathbf{M}\ddot{\mathbf{x}} + \mathbf{C}\dot{\mathbf{x}} + \mathbf{K}\mathbf{x} = \mathbf{f}$. By observation, we find the following:

1. The mass, damping, and stiffness matrices are symmetric.
2. All diagonal entries of these matrices are non-negative.
3. The off-diagonal elements of both the damping and the stiffness matrices are non-positive.
4. The off-diagonal elements of the mass matrix are non-negative.

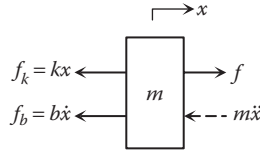
These results are true for stable mechanical systems with purely translational or rotational motion. The reader can use them as necessary conditions to check the correctness of the differential equations of motion.

5.2.6 D'Alembert's Principle

Newton's second law can be reformulated as

$$\sum \mathbf{F} - m\mathbf{a}_C = 0 \quad (5.24)$$

which is known as D'Alembert's principle. The equation has $\sum \mathbf{F}$ as the sum of all the physical forces and $-m\mathbf{a}_C$ as the inertial force, which is a fictitious force. The minus sign associated with the inertial force indicates that the force acts in the negative direction when $\mathbf{a}_C > 0$. If the inertial force is included with the external forces, the mass can be considered to be in equilibrium. D'Alembert's principle is completely equivalent to the formulation of Newton's second law, although it looks like a classical static force balance. To use D'Alembert's principle correctly, the inertial force must be shown in the free-body diagrams correctly. The following is a simple example that shows the derivation of the differential equation of motion by using D'Alembert's principle.

**FIGURE 5.37**

Free-body diagram of a mass–spring–damper system by using D'Alembert's principle.

Example 5.7: D'Alembert's Principle

Reconsider the single-degree-of-freedom mass–spring–damper system in Example 5.4. Draw a free-body diagram and derive the differential equation of motion by using D'Alembert's principle.

Solution

The displacement x is chosen as the generalized coordinate, and the origin is set at the static equilibrium position. For D'Alembert's principle, the free-body diagram of the mass is shown in Figure 5.37, where the inertial force is shown by a dashed line. Note that we assumed $x > 0$ and $\ddot{x} > 0$. The inertial force acts in the negative direction with a magnitude of $m\ddot{x}$. Applying D'Alembert's principle in the x direction results in

$$+\rightarrow x: \sum F_x - ma_x = 0$$

$$f - kx - b\dot{x} - m\ddot{x} = 0$$

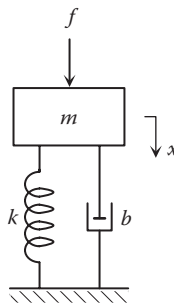
or

$$m\ddot{x} + b\dot{x} + kx = f$$

which is the same as the one obtained previously in Example 5.4.

PROBLEM SET 5.2

1. For the system shown in Figure 5.38, the input is the force f and the output is the displacement x of the mass.
 - a. Draw the necessary free-body diagram and derive the differential equation of motion.

**FIGURE 5.38**

Problem 1.

- b. Using the differential equation obtained in Part (a), determine the transfer function. Assume that initial conditions are $x(0) = 0$ and $\dot{x}(0) = 0$.
 - c. Using the differential equation obtained in Part (a), determine the state-space representation.
2. Repeat Problem 1 for the system shown in Figure 5.39.
3. Repeat Problem 1 for the system shown in Figure 5.40.
4. Repeat Problem 1 for the system shown in Figure 5.41.
5.  For Problems 1 through 4, use Simulink to construct block diagrams to find the displacement output $x(t)$ of the system subjected to an applied force $f(t) = 10u(t)$, where $u(t)$ is the unit-step function. The parameter values are $m = 1$ kg, $b = 2$ N·s/m, and $k = 5$ N/m. Assume zero initial conditions.

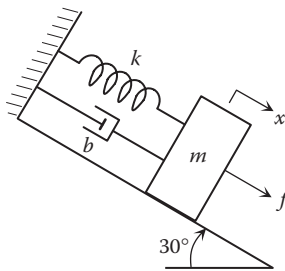


FIGURE 5.39
Problem 2.

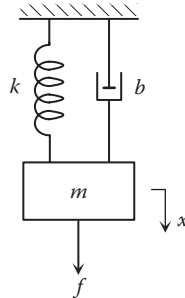


FIGURE 5.40
Problem 3.

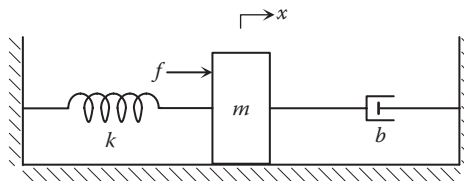


FIGURE 5.41
Problem 4.

6. The system shown in Figure 5.42 simulates a machine supported by rubbers, which are approximated as four identical spring–damper units. The input is the force f and the output is the displacement x of the mass. The parameter values are $m = 500$ kg, $b = 250$ N·s/m, and $k = 200,000$ N/m.
- Draw the necessary free-body diagram and derive the differential equation of motion.
 - Determine the transfer function. Assume zero initial conditions.
 - Determine the state-space representation.
 - Find the transfer function from the state-space form and compare with the result obtained in Part (b).
7. Repeat Problem 6 for the system shown in Figure 5.43.
8. The system shown in Figure 5.44 simulates a vehicle traveling on a rough road. The input is the displacement z .
- Draw the necessary free-body diagram and derive the differential equation of motion.

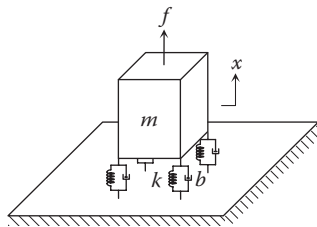


FIGURE 5.42
Problem 6.

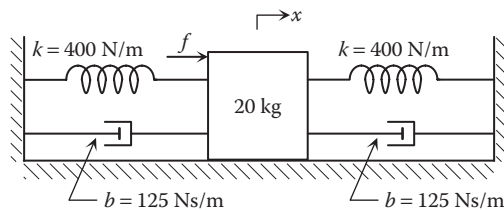


FIGURE 5.43
Problem 7.

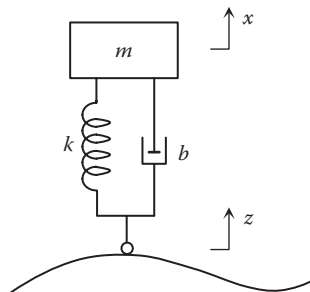


FIGURE 5.44
Problem 8.

- b. Assuming zero initial conditions, determine the transfer function for two different cases of output: (1) displacement x and (2) velocity $\dot{x}$.
 - c. Determine the state-space representation for two different cases of output: (1) displacement x and (2) velocity $\dot{x}$.
9. Repeat Problem 8 for the system shown in Figure 5.45, where the cam and follower impart a displacement z to the lower end of the system.
10. For the system shown in Figure 5.46, the input is the force f and the outputs are the displacements x_1 and x_2 of the masses.
 - a. Draw the necessary free-body diagrams and derive the differential equations of motion.
 - b. Write the differential equations of motion in the second-order matrix form.
 - c. Using the differential equations obtained in Part (a), determine the state-space representation.

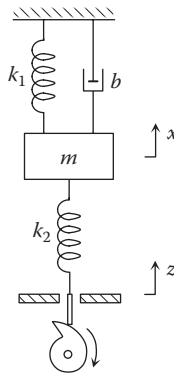


FIGURE 5.45
Problem 9.

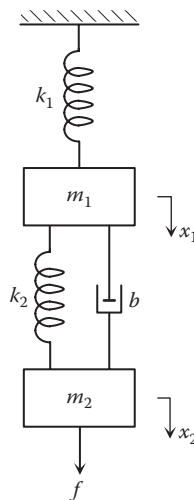


FIGURE 5.46
Problem 10.

11. Repeat Problem 10 for the system shown in Figure 5.47.
12. Repeat Problem 10 for the system shown in Figure 5.48.
13.  For Problems 10 through 12, use MATLAB commands to define the systems in the state-space form and then convert to the transfer function form. Assume that the displacements of the two masses, x_1 and x_2 , are the outputs, and all initial conditions are zero. The masses are $m_1 = 5$ kg and $m_2 = 15$ kg. The spring constants are $k_1 = 7.5$ kN/m and $k_2 = 15$ kN/m. The viscous damping coefficients are $b_1 = b_2 = b = 250$ N·s/m.
14. For the system in Figure 5.49, the inputs are the forces f_1 and f_2 applied to the masses and the outputs are the displacements x_1 and x_2 of the masses.
 - a. Draw the necessary free-body diagrams and derive the differential equations of motion.
 - b. Write the differential equations of motion in the second-order matrix form.
 - c. Using the differential equations obtained in Part (a), determine the state-space representation.

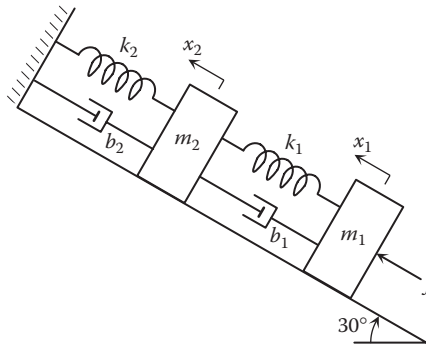


FIGURE 5.47
Problem 11.

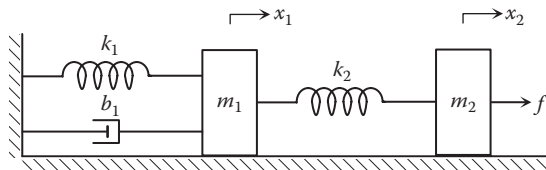


FIGURE 5.48
Problem 12.

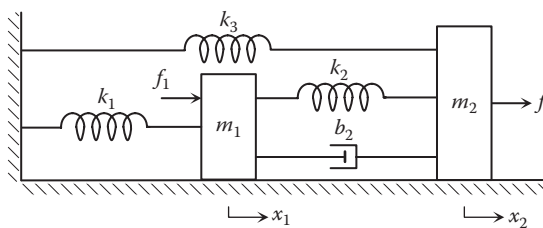


FIGURE 5.49
Problem 14.

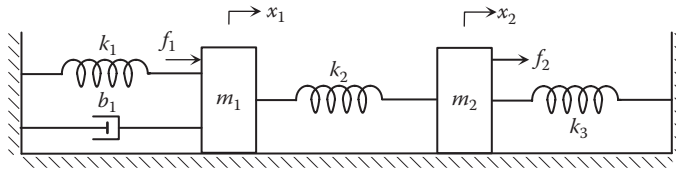


FIGURE 5.50
Problem 15.

15. Repeat Problem 14 for the system shown in [Figure 5.50](#).
16.  For Problems 14 and 15, use MATLAB commands to define the systems in the state-space form and then convert to the transfer function form. Assume that the displacements of the two masses, x_1 and x_2 , are the outputs, and all initial conditions are zero. The masses are $m_1 = 5$ kg and $m_2 = 15$ kg. The spring constants are $k_1 = 7.5$ kN/m, $k_2 = 15$ kN/m, and $k_3 = 30$ kN/m. The viscous damping coefficients are $b_1 = b_2 = 250$ N·s/m.
17. For the system in [Figure 5.51](#), the input is the force f and the outputs are the displacement x_1 of the mass and the displacement x_2 of the massless junction A.
 - a. Draw the necessary free-body diagrams and derive the differential equations of motion. Determine the number of degrees of freedom and the order of the system.
 - b. Write the differential equations of motion in the second-order matrix form.
 - c. Using the differential equation obtained in Part (a), determine the state-space representation.
18. Repeat Problem 17 for the system in [Figure 5.52](#), where the input is the displacement z and the outputs are the displacements x_1 and x_2 .

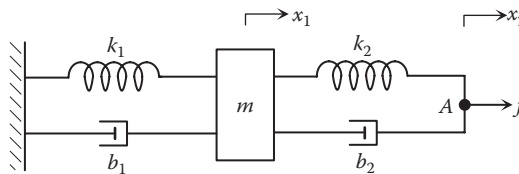


FIGURE 5.51
Problem 17.

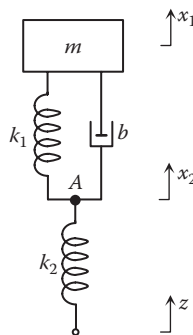


FIGURE 5.52
Problem 18.

5.3 Rotational Systems

In this section, we consider the derivation of a mathematical model for a rotational mechanical system. When a rigid body moves arbitrarily in three-dimensional space, the axis of rotation keeps changing. This makes modeling rather complex. Thus, the discussion of three-dimensional rigid body motions is not covered in this text, and we are mainly concerned with two-dimensional or plane motion. The fundamentals of rigid bodies in three dimensions are included only to help the interested reader in further studies.

5.3.1 General Moment Equation

The moment equation is applicable to systems of particles and rigid bodies in three and two dimensions. For a system of particles connected rigidly or a rigid body in arbitrary motion, using Newton's second law leads to the general moment equation given by

$$\sum \mathbf{M}_P = \dot{\mathbf{H}}_P + m\mathbf{r}_{C/P} \times \mathbf{a}_P \quad (5.25)$$

where:

- subscript P denotes an arbitrary accelerating point
- subscript C denotes the mass center of the system of particles or the rigid body
- $\sum \mathbf{M}_P$ is the sum of all externally applied moments about point P
- $\mathbf{H}_P$ is the angular momentum vector about point P
- $\dot{\mathbf{H}}_P$ is the time rate of change of $\mathbf{H}_P$
- m is the total mass of the system of particles or the mass of the rigid body
- $\mathbf{r}_{C/P}$ is the position vector of the mass center C with respect to point P
- $\mathbf{a}_P$ is the acceleration vector of point P

It is usually difficult to obtain $\dot{\mathbf{H}}_P$ for complex systems. For a rigid body, the angular momentum is related to mass moments of inertia and the angular velocity,

$$\mathbf{H}_P = \mathbf{I}_P \boldsymbol{\omega} \quad (5.26)$$

or

$$\begin{Bmatrix} H_x \\ H_y \\ H_z \end{Bmatrix}_P = \begin{bmatrix} I_{xx} & I_{xy} & I_{xz} \\ I_{yx} & I_{yy} & I_{yz} \\ I_{zx} & I_{zy} & I_{zz} \end{bmatrix}_P \begin{Bmatrix} \omega_x \\ \omega_y \\ \omega_z \end{Bmatrix} \quad (5.27)$$

where:

- $\mathbf{H}_P$ is the angular momentum about point P
- $\boldsymbol{\omega}$ is the angular velocity vector (3×1)
- $\mathbf{I}_P$ is the mass moment of inertia tensor (3×3) about point P, and

$$I_{xx} = \int (y^2 + z^2) dm, \quad I_{yy} = \int (x^2 + z^2) dm, \quad I_{zz} = \int (x^2 + y^2) dm \quad (5.28)$$

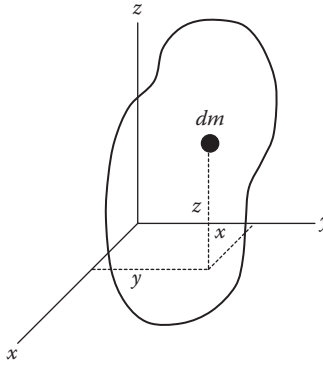


FIGURE 5.53
A differential element of a rigid body.

$$I_{xy} = I_{yx} = -\int xy \, dm, \quad I_{yz} = I_{zy} = -\int yz \, dm, \quad I_{xz} = I_{zx} = -\int xz \, dm \quad (5.29)$$

As shown in Figure 5.53, the integrals containing squares represent the mass moments of inertia of the body about the x , y , and z axes, respectively. The integrals containing products of coordinates represent the mass products of inertia of the body.

In general, the time derivatives of the nine elements in the matrix $\mathbf{I}_P$ are non-zero for a nonsymmetric rigid body, and thus, it is difficult to study the dynamics of a rigid body in three dimensions. However, if the motion of the rigid body is restricted to a plane, the complexity is reduced significantly.

5.3.2 Modeling of Rigid Bodies in Plane Motion

In many engineering applications, the motion of rigid bodies is primarily in two dimensions. A rigid body is in plane motion if it translates in two dimensions (or a plane) and rotates only about an axis that is perpendicular to the plane. For a rigid body in plane motion, the mass moment of inertia is a scalar quantity and the time rate of change of the angular momentum reduces to

$$\dot{\mathbf{H}}_P = I_P \boldsymbol{\alpha} \quad (5.30)$$

where:

I_P is the mass moment of inertia of the body about an axis through point P

$\boldsymbol{\alpha}$ is the angular acceleration

Then, Equation 5.25 becomes

$$\sum \mathbf{M}_P = I_P \boldsymbol{\alpha} + m \mathbf{r}_{C/P} \times \mathbf{a}_P \quad (5.31)$$

Note that the rigid body in plane motion is constrained to rotate about only one axis. Therefore, the net moment $\Sigma \mathbf{M}_P$ and the angular acceleration α are essentially scalars, whose signs signify the directions (e.g., clockwise or counterclockwise). Because the cross-product term is $m\mathbf{r}_{C/P} \times \mathbf{a}_P = \mathbf{r}_{C/P} \times (m\mathbf{a}_P)$, it can be considered as the effective moment caused by a fictitious force $m\mathbf{a}_P$, and the direction of the effective moment can be denoted by its sign.

If the rigid body rotates about a fixed axis through point O, Equation 5.31 can be simplified as

$$\sum M_O = I_O \alpha \quad (5.32)$$

with $P = O$ and $a_P = a_O = 0$. I_O is the mass moment of inertia of the body about point O. If the axis of rotation is not fixed, the dynamics model of the rigid body can be derived using Equation 5.31 or

$$\sum M_C = I_C \alpha \quad (5.33)$$

with $P = C$ and $r_{C/P} = 0$. I_C is the mass moment of inertia of the body about the mass center C. Both Equations 5.31 and 5.33 are applicable regardless of whether the axis of rotation is fixed or not.

Example 5.8: A Single-Degree-of-Freedom Rotational Mass–Spring–Damper System

Consider a simple disk–shaft system shown in Figure 5.54a, in which the disk rotates about a fixed axis through point O. A single-degree-of-freedom torsional mass–spring–damper system in Figure 5.54b can be used to approximate the dynamic behavior of the disk–shaft system. I_O is the mass moment of inertia of the disk about point O, K represents the elasticity of the shaft, and B represents torsional viscous damping. Derive the differential equation of motion.

Solution

The free-body diagram of the disk is shown in Figure 5.54c. Because the disk rotates about a fixed axis, we can apply Equation 5.32 about the fixed point O. Assuming that counterclockwise is the positive direction, we have

$$\begin{aligned} +\curvearrowright : \sum M_O &= I_O \alpha \\ \tau - K\theta - B\dot{\theta} &= I_O \ddot{\theta} \end{aligned}$$

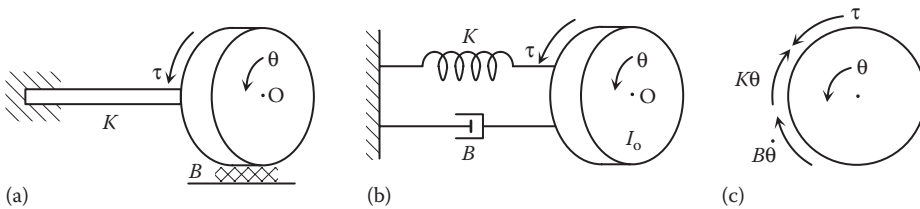


FIGURE 5.54

A rotational mass–spring–damper system: (a) physical system, (b) mass–spring–damper model, and (c) free-body diagram.

Thus,

$$I_O \ddot{\theta} + B \dot{\theta} + K \theta = \tau$$

Example 5.9: A Two-Degree-of-Freedom Rotational Mass–Spring–Damper System

Consider the disk–shaft system shown in [Figure 5.55](#). The mass moments of inertia of the two disks about their longitudinal axes are I_1 and I_2 , respectively. The massless torsional springs represent the elasticity of the shafts and the torsional viscous dampers represent the fluid coupling.

- Draw the necessary free-body diagrams and derive the differential equations of motion.
- Determine the transfer functions $\Theta_1(s)/T(s)$ and $\Theta_2(s)/T(s)$. All initial conditions are assumed to be zero.

Solution

- We choose the angular displacements θ_1 and θ_2 as the generalized coordinates. Assume that $\theta_1 > \theta_2 > 0$. The free-body diagrams are shown in [Figure 5.56](#). Applying Equation 5.32 about the fixed points O_1 and O_2 , respectively, gives

$$\begin{aligned}
 +\curvearrowright: \quad \sum M_O &= I_O \alpha \\
 \tau - K_1 \theta_1 - B_1 \dot{\theta}_1 - K_3(\theta_1 - \theta_2) &= I_1 \ddot{\theta}_1 \\
 K_3(\theta_1 - \theta_2) - K_2 \theta_2 - B_2 \dot{\theta}_2 &= I_2 \ddot{\theta}_2
 \end{aligned}$$

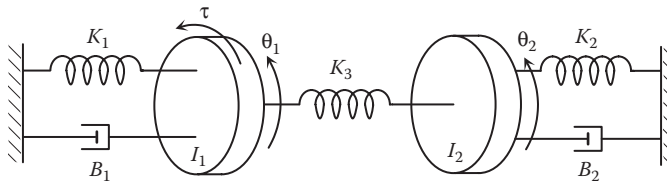


FIGURE 5.55

A two-degree-of-freedom rotational disk–shaft system.

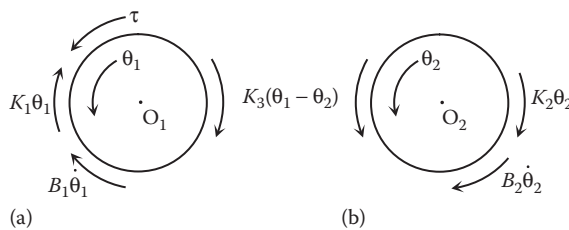


FIGURE 5.56

Free-body diagrams for the disk–shaft system in [Figure 5.55](#).

Rearranging the equations results in

$$I_1\ddot{\theta}_1 + B_1\dot{\theta}_1 + (K_1 + K_3)\theta_1 - K_3\theta_2 = \tau$$

$$I_2\ddot{\theta}_2 + B_2\dot{\theta}_2 - K_3\theta_1 + (K_2 + K_3)\theta_2 = 0$$

b. Taking Laplace transform gives

$$I_1s^2\Theta_1(s) + B_1s\Theta_1(s) + (K_1 + K_3)\Theta_1(s) - K_3\Theta_2(s) = T(s)$$

$$I_2s^2\Theta_2(s) + B_2s\Theta_2(s) - K_3\Theta_1(s) + (K_2 + K_3)\Theta_2(s) = 0$$

or in matrix form

$$\begin{bmatrix} I_1s^2 + B_1s + K_1 + K_3 & -K_3 \\ -K_3 & I_2s^2 + B_2s + K_2 + K_3 \end{bmatrix} \begin{Bmatrix} \Theta_1(s) \\ \Theta_2(s) \end{Bmatrix} = \begin{Bmatrix} T(s) \\ 0 \end{Bmatrix}$$

Using Cramer's rule, we can solve for $\Theta_1(s)/T(s)$ and $\Theta_2(s)/T(s)$ as

$$\frac{\Theta_1(s)}{T(s)} = \frac{I_2s^2 + B_2s + K_2 + K_3}{(I_1s^2 + B_1s + K_1 + K_3)(I_2s^2 + B_2s + K_2 + K_3) - K_3^2}$$

$$\frac{\Theta_2(s)}{T(s)} = \frac{K_3}{(I_1s^2 + B_1s + K_1 + K_3)(I_2s^2 + B_2s + K_2 + K_3) - K_3^2}$$

5.3.3 Mass Moment of Inertia

As shown in Equation 5.28, the mass moment of inertia of a rigid body about a specified axis of rotation is defined as

$$I = \int r^2 dm \quad (5.34)$$

where r is the distance between the axis of rotation and the mass element dm . The mass moments of inertia for some rigid bodies with common shapes are given in Table 5.1, where all masses are assumed to be uniformly distributed and the axes of rotation all pass through the mass centers. If the axis of rotation does not coincide with the axis through the mass center, but is parallel to it, the parallel-axis theorem can be applied to obtain the corresponding moment of inertia,

$$I = I_C + md^2 \quad (5.35)$$

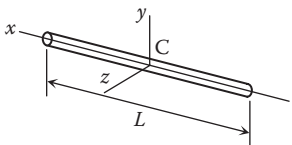
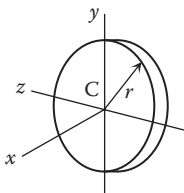
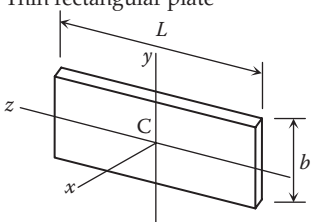
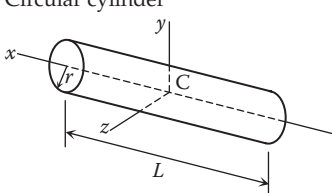
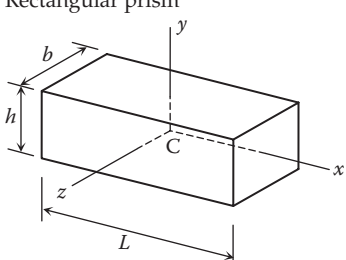
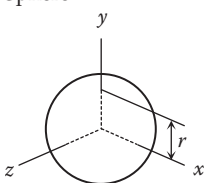
where:

I_C is the mass moment of inertia of the rigid body about an axis passing through its center of mass

I is the mass moment of inertia about the new axis

d is the perpendicular distance between the two parallel axes

TABLE 5.1
Mass Moments of Inertia of Common Geometric Shapes

Slender rod 	$I_y = I_z = \frac{1}{12}mL^2$ $I_x = 0$
Thin disk 	$I_x = \frac{1}{2}mr^2$ $I_y = I_z = \frac{1}{4}mr^2$
Thin rectangular plate 	$I_x = \frac{1}{12}m(L^2 + b^2)$ $I_y = \frac{1}{12}mL^2$ $I_z = \frac{1}{12}mb^2$
Circular cylinder 	$I_x = \frac{1}{2}mr^2$ $I_y = I_z = \frac{1}{12}m(3r^2 + L^2)$
Rectangular prism 	$I_x = \frac{1}{12}m(b^2 + h^2)$ $I_y = \frac{1}{12}m(L^2 + b^2)$ $I_z = \frac{1}{12}m(L^2 + h^2)$
Sphere 	$I_x = I_y = I_z = \frac{2}{5}mr^2$

Source: Beer, F.P. et al., *Vector Mechanics for Engineers: Dynamics*, 9th ed., McGraw-Hill, New York, NY, 2009; Hibbeler, R.C., *Engineering Mechanics: Dynamics*, 12th ed., Prentice Hall, Upper Saddle River, NJ, 2010. With Permission.

Example 5.10: An Inverted-Pendulum–Bob System

Consider the inverted-pendulum system shown in Figure 5.57a, in which a point mass m is attached at the tip of a uniform slender rod of mass M and length L . The mass center of the rod is located at point C . The inverted-pendulum system rotates about an axis through the joint O . The friction at the joint is modeled as a torsional viscous damper of coefficient B .

- Determine the mass moment of inertia of the system about point O .
- Draw the free-body diagram for the inverted-pendulum system and obtain the nonlinear equation of motion.
- Linearize the equation of motion for small angles θ .

Solution

- The system consists of a point mass and a slender rod. The total mass moment of inertia about point O is

$$I_O = I_{O_mass} + I_{O_rod}$$

where

$$I_{O_mass} = mL^2$$

and I_{O_rod} can be obtained using the parallel-axis theorem,

$$I_{O_rod} = I_{C_rod} + Md^2 = \frac{1}{12}ML^2 + M\left(\frac{L}{2}\right)^2 = \frac{1}{3}ML^2$$

Then,

$$I_O = \left(m + \frac{M}{3}\right)L^2$$

- For the inverted-pendulum system, the free-body diagram is shown in Figure 5.57b, where R_x and R_y are the x and y components of the reaction force at the joint O , respectively. Note that the system rotates about a fixed axis through point O . Applying Equation 5.32 about the fixed point O gives

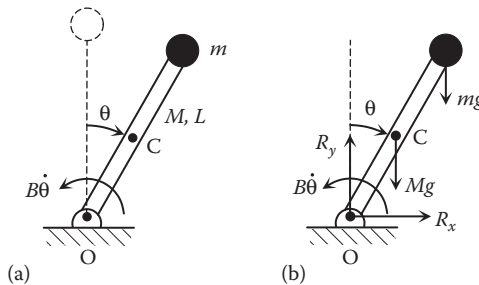


FIGURE 5.57

An inverted pendulum-bob system: (a) physical system and (b) free-body diagram.

$$+\curvearrowright: \sum M_O = I_O \alpha$$

$$\frac{1}{2} L \sin \theta \cdot Mg + L \sin \theta \cdot mg - B \dot{\theta} = I_O \ddot{\theta}$$

Substituting I_O obtained in Part (a) into the equation, and rearranging it in the input–output differential equation form, we obtain

$$\left(m + \frac{M}{3}\right) L^2 \ddot{\theta} + B \dot{\theta} - \left(m + \frac{M}{2}\right) g L \sin \theta = 0$$

c. For small angles θ , $\sin \theta \approx \theta$. The equation of motion becomes

$$\left(m + \frac{M}{3}\right) L^2 \ddot{\theta} + B \dot{\theta} - \left(m + \frac{M}{2}\right) g L \theta = 0$$

which is a linear equation in terms of θ .

Let us examine the nonlinear differential equation of motion in Part (b). Note that the unknown reaction forces R_x and R_y at the joint O do not appear in the moment equation $\Sigma M_O = I_O \alpha$. If an arbitrary nonfixed point P is used, we must apply Equation 5.31. The moments caused by the unknown forces R_x and R_y appear in the equation, and auxiliary equations are required to eliminate the terms related to R_x and R_y . This is the advantage of choosing the fixed point to apply the moment equation if a rigid body rotates about a fixed axis.

Example 5.11: A Coupled Pendulum System

Consider the two-degree-of-freedom system shown in [Figure 5.58a](#), where two uniform slender rods of mass m and length L are connected by a translational spring of stiffness k . θ_1 and θ_2 are the angular displacements of the rods. The torques τ_1 and τ_2 are applied to the pivot points O_1 and O_2 , respectively. When $\theta_1 = 0$, $\theta_2 = 0$, $\tau_1 = 0$, and $\tau_2 = 0$, the spring is at its free length. Draw the necessary free-body diagrams and derive the differential equations of motion. Assume small angles for θ_1 and θ_2 .

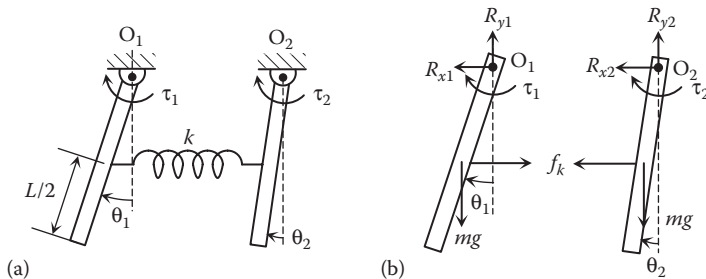


FIGURE 5.58

A coupled pendulum system: (a) physical system and (b) free-body diagram.

Solution

We choose the angular displacements θ_1 and θ_2 as the generalized coordinates. Assume $\theta_1 > \theta_2 > 0$, which implies that the spring is in tension. The free-body diagrams are shown in Figure 5.58b. Applying Equation 5.32 about the fixed points O_1 and O_2 , respectively, gives

$$\begin{aligned}
 + \curvearrowright: \quad \sum M_O &= I_O \alpha \\
 \tau_1 - \frac{1}{2} L \sin \theta_1 \cdot mg - \frac{1}{2} L \cos \theta_1 \cdot f_k &= I_{O_1} \ddot{\theta}_1 \\
 \tau_2 - \frac{1}{2} L \sin \theta_2 \cdot mg + \frac{1}{2} L \cos \theta_2 \cdot f_k &= I_{O_2} \ddot{\theta}_2
 \end{aligned}$$

where $I_{O_1} = I_{O_2} = \frac{1}{3} mL^2$. Note that the spring force f_k is in the horizontal direction due to the small-angle assumption, and its magnitude can be approximated as $k(\frac{1}{2}L\sin\theta_1 - \frac{1}{2}L\sin\theta_2)$. Substituting f_k into the equations and rearranging them results in

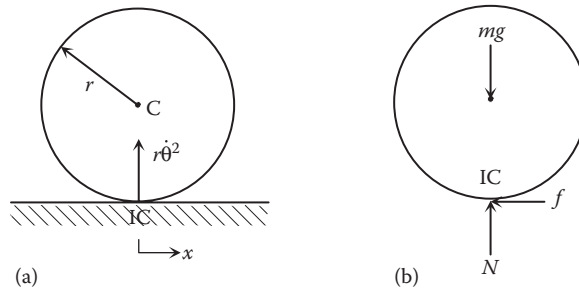
$$\begin{aligned}
 \frac{1}{3} mL^2 \ddot{\theta}_1 + \frac{1}{4} kL^2 \cos \theta_1 (\sin \theta_1 - \sin \theta_2) + \frac{1}{2} mgL \sin \theta_1 &= \tau_1 \\
 \frac{1}{3} mL^2 \ddot{\theta}_2 - \frac{1}{4} kL^2 \cos \theta_2 (\sin \theta_1 - \sin \theta_2) + \frac{1}{2} mgL \sin \theta_2 &= \tau_2
 \end{aligned}$$

For small angles, we have

$$\begin{aligned}
 \frac{1}{3} mL^2 \ddot{\theta}_1 + \frac{1}{4} kL^2 (\theta_1 - \theta_2) + \frac{1}{2} mgL \theta_1 &= \tau_1 \\
 \frac{1}{3} mL^2 \ddot{\theta}_2 - \frac{1}{4} kL^2 (\theta_1 - \theta_2) + \frac{1}{2} mgL \theta_2 &= \tau_2
 \end{aligned}$$

5.3.4 Pure Rolling Motion

Wheels are common mechanical systems involving general plane motion. Figure 5.59a shows a uniform disk rolling on a horizontal surface. If there is no slipping between the disk and the surface, the disk undergoes pure rolling motion. The contact point is the instantaneous center (IC), where the velocity is zero. From the results of kinematics, the acceleration of the IC is $r\dot{\theta}^2$, and its direction points from the IC to the mass center C (or for a uniform disk, the geometric center). The problem of finding a_{IC} is left to the reader as an exercise.

**FIGURE 5.59**

A pure rolling disk: (a) physical system and (b) free-body diagram.

The free-body diagram of the disk is shown in Figure 5.59b, where reaction forces at the contact point include the normal force N and the friction force f . Note that a fixed point does not exist for the rolling motion; thus, we cannot use the moment equation $\Sigma M_O = I_O \alpha$. Also, it is inconvenient to sum the moments about the mass center by using the equation $\Sigma M_C = I_C \alpha$, because the unknown reaction force f inevitably appears in the equation and needs to be eliminated with the help of auxiliary equations. However, if the point IC is used, the moment equation

$$\Sigma \mathbf{M}_{IC} = I_{IC} \alpha + m \mathbf{r}_{C/IC} \times \mathbf{a}_{IC} \quad (5.36)$$

reduces to

$$\Sigma M_{IC} = I_{IC} \alpha. \quad (5.37)$$

Because the position vector $\mathbf{r}_{C/IC}$ and the acceleration vector $\mathbf{a}_{IC}$ are parallel, their cross product is zero, $m \mathbf{r}_{C/IC} \times \mathbf{a}_{IC} = 0$. Therefore, we do not need to deal with the moment caused by the friction force at the contact point. This is a special way of using Equation 5.31.

In summary, if there is a point S such that the position vector $\mathbf{r}_{C/S}$ is parallel to the acceleration vector $\mathbf{a}_S$, we can apply

$$\Sigma M_S = I_S \alpha \quad (5.38)$$

to model the dynamics of the rigid body. I_S the mass moment of inertia of the body about point S. For the pure rolling disk in Figure 5.59, the IC at the contact point has the property of $\mathbf{r}_{C/IC} \parallel \mathbf{a}_{IC}$.

Example 5.12: A Pure Rolling Disk

Consider the system shown in Figure 5.60a, in which a uniform disk of mass m and radius r rolls on a horizontal surface. A translational spring of stiffness k is attached to the disk. Assuming that there is no slipping between the disk and the surface, derive the differential equation of motion.

Solution

The free-body diagram of the system is shown in Figure 5.60b, in which the normal force N and the friction force f are reaction forces at the contact point. When the disk is at the static equilibrium position, we have $f_k = k \delta_{st}$, where δ_{st} is the static deformation of the spring. Then,

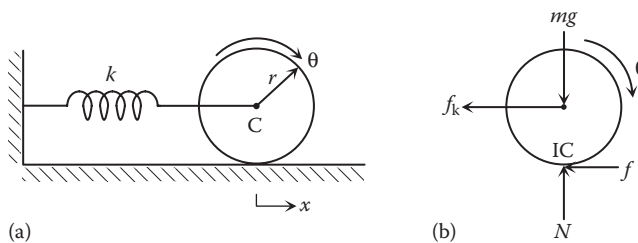


FIGURE 5.60

A pure rolling disk-spring system: (a) physical system and (b) free-body diagram.

$$+ \curvearrowright: \sum M_{IC} = 0$$

$$r \cdot k \delta_{st} = 0$$

or $\delta_{st} = 0$. We choose the static equilibrium position as the origin. When the disk rolls, the spring force is $f_k = k(x + \delta_{st}) = kx$. Because of no slipping, the contact point is the IC, and $\mathbf{r}_{C/IC} \parallel \mathbf{a}_{IC}$. Applying Equation 5.37 gives

$$+ \curvearrowright: \sum M_{IC} = I_{IC} \alpha$$

$$-r \cdot kx = I_{IC} \alpha$$

where

$$I_{IC} = I_C + md^2 = \frac{1}{2}mr^2 + mr^2 = \frac{3}{2}mr^2$$

Introducing the assumption of no slipping, $x = r\theta$, we obtain the differential equation of motion

$$\frac{3}{2}mr^2\ddot{\theta} + kr^2\theta = 0$$

Strictly speaking, the disk in [Figure 5.59](#) or Example 5.12 is not a purely rotational system, and the rolling motion involves both translation and rotation. Such a system can also be modeled using both Newton's second law and the moment equation. We will present the corresponding discussion in the next section.

PROBLEM SET 5.3

1. Consider the rotational system shown in [Figure 5.61](#). The system consists of a massless shaft and a uniform thin disk of mass m and radius r . The disk is constrained to rotate about a fixed longitudinal axis along the shaft. The shaft is equivalent to a torsional spring of stiffness K connected in parallel with a torsional viscous damper of damping coefficient B . Draw the necessary free-body diagram and derive the differential equation of motion.
2. Repeat Problem 1 for the system shown in [Figure 5.62](#).

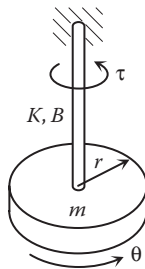


FIGURE 5.61
Problem 1.

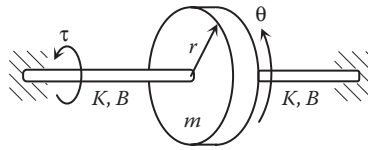


FIGURE 5.62
Problem 2.

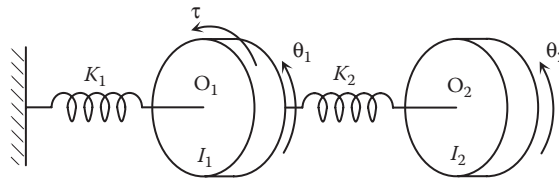


FIGURE 5.63
Problem 3.

3. Consider the torsional mass–spring system in [Figure 5.63](#). The mass moments of inertia of the two disks are I_1 and I_2 , respectively. The massless torsional springs represent the elasticity of the shafts.
 - a. Draw the necessary free-body diagrams and derive the differential equations of motion. Provide the equations in the second-order matrix form.
 - b. Determine the transfer functions $\Theta_1(s)/T(s)$ and $\Theta_2(s)/T(s)$. All the initial conditions are assumed to be zero.
 - c. Determine the state-space representation with the angular displacements θ_1 and θ_2 as the outputs.
4. Repeat Problem 3 for the system shown in [Figure 5.64](#), where the torsional viscous damper represents the fluid coupling. The input is the angular displacement ϕ at the end of the shaft.
5. Consider the pendulum system shown in [Figure 5.65](#). The system consists of a bob of mass m and a uniform rod of mass M and length L . The pendulum pivots at the joint O. Draw the necessary free-body diagram and derive the differential equation of motion. Assume small angles for θ .
6. Repeat Problem 5 for the systems shown in [Figure 5.66](#), in which (a) the mass of the rod is neglected and (b) no bob is attached to the rod.

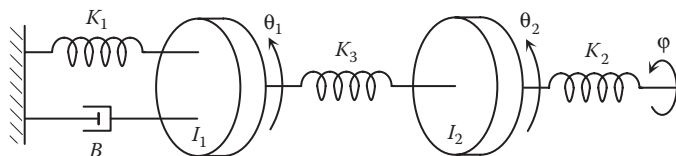


FIGURE 5.64
Problem 4.

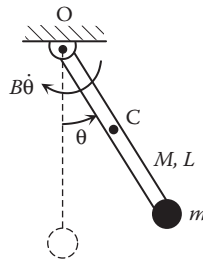


FIGURE 5.65
Problem 5.

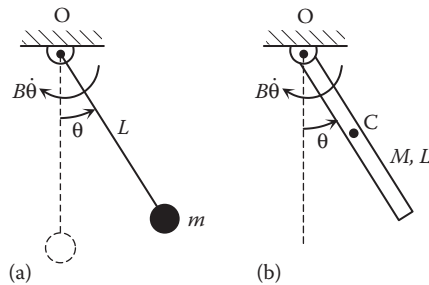


FIGURE 5.66
Problem 6.

7. The system shown in Figure 5.67 consists of a uniform rod of mass M and length L and a translational spring of stiffness k at the rod's left tip. The friction at the joint O is modeled as a damper with coefficient of torsional viscous damping B . The input is the force f and the output is the angle θ . The position $\theta = 0$ corresponds to the static equilibrium position when $f = 0$.
- Draw the necessary free-body diagram and derive the differential equation of motion for small angles θ .
 - Using the linearized differential equation obtained in Part (a), determine the transfer function $\Theta(s)/F(s)$. Assume that the initial conditions are $\theta(0) = 0$ and $\dot{\theta}(0) = 0$.
 - Using the differential equation obtained in Part (a), determine the state-space representation.

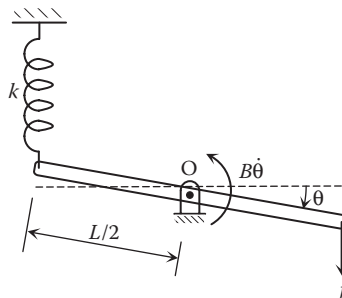


FIGURE 5.67
Problem 7.

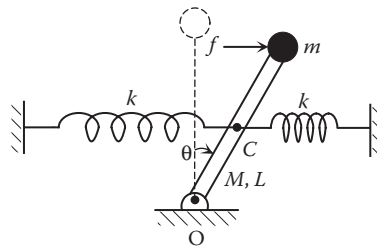


FIGURE 5.68
Problem 8.

8. Repeat Problem 7 for the system shown in Figure 5.68. When $\theta = 0$ and $f = 0$, the springs are at their free lengths.
9. Example 5.4 Part (d) shows how one can represent a linear system in Simulink based on the differential equation of the system. A linear system can also be represented in transfer function or state-space form. The corresponding blocks in Simulink are Transfer Fcn and State-Space, respectively. Consider Problem 7 and construct a Simulink block diagram to find the output $\theta(t)$ of the system, which is represented using (a) the linearized differential equation of motion, (b) the transfer function, and (c) the state-space form obtained in Problem 7. The parameter values are $M = 0.8$ kg, $L = 0.6$ m, $k = 100$ N/m, $B = 0.4$ N·s/m, and $g = 9.81$ m/s². The input force f is the unit-impulse function, which has a magnitude of 20 N and a time duration of 0.01 s.
10. Repeat Problem 9 by using the linearized differential equation of motion, the transfer function, and the state-space form obtained in Problem 8. The parameter values are $m = 0.2$ kg, $M = 0.8$ kg, $L = 0.6$ m, $k = 100$ N/m, and $g = 9.81$ m/s². The input force f is the unit-impulse function, which has a magnitude of 20 N and a time duration of 0.01 s.
11. Consider the system shown in Figure 5.69, in which the motion of the rod is small angular rotation. When $\theta = 0$, the springs are at their free lengths.
 - a. Determine the mass moment of inertia of the rod about point O. Assume that $a > b$.
 - b. Draw the necessary free-body diagram and derive the differential equation of motion for small angles θ .

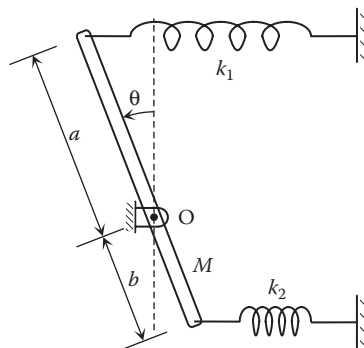
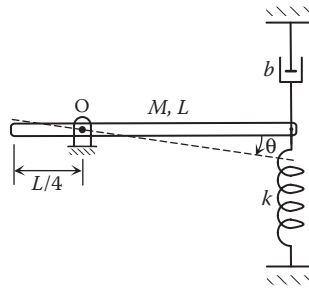
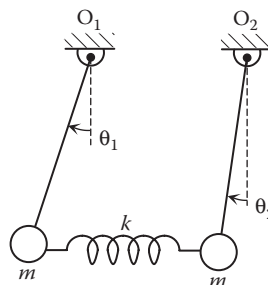


FIGURE 5.69
Problem 11.

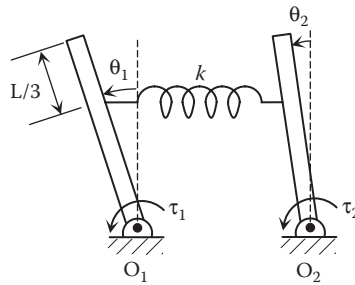
**FIGURE 5.70**

Problem 12.

12. Consider the system shown in Figure 5.70, in which a lever arm has a spring-damper combination on the other side. When $\theta = 0$, the system is in static equilibrium.
 - a. Assuming that the lever arm can be approximated as a uniform slender rod, determine the mass moment of inertia of the rod about point O.
 - b. Draw the necessary free-body diagram and derive the differential equation of motion for small angles θ .
13. Consider the two-degree-of-freedom system shown in Figure 5.71, in which two simple inverted pendulums are connected by a translational spring of stiffness k . Each pendulum consists of a point mass m concentrated at the tip of a massless rope of length L . θ_1 and θ_2 are the angular displacements of the pendulums. When $\theta_1 = 0$ and $\theta_2 = 0$, the spring is at its free length.
 - a. Draw the necessary free-body diagrams and derive the differential equations of motion. Assume small angles for θ_1 and θ_2 . Provide the equations in the second-order matrix form.
 - b. Using the differential equations obtained in Part (a), determine the state-space representation with the angular velocities $\dot{\theta}_1$ and $\dot{\theta}_2$ as the outputs.
14. Repeat Problem 13 for the system shown in Figure 5.72, in which each pendulum is a uniform slender rod of mass m and length L . The inputs are the torques τ_1 and τ_2 applied to the pivots O_1 and O_2 , respectively. When $\theta_1 = 0$, $\theta_2 = 0$, $\tau_1 = 0$, and $\tau_2 = 0$, the spring is at its free length.

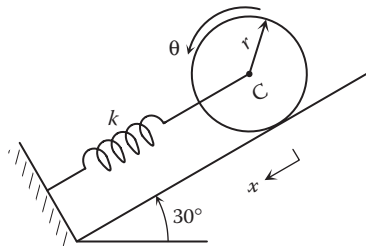
**FIGURE 5.71**

Problem 13.

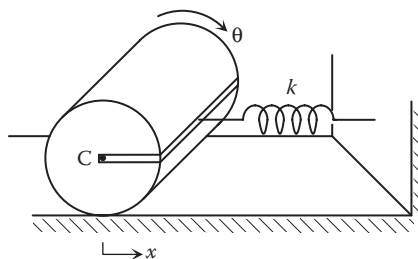
**FIGURE 5.72**

Problem 14.

15. Consider the system shown in Figure 5.73, in which a uniform sphere of mass m and radius r rolls along an inclined plane of 30° . A translational spring of stiffness k is attached to the sphere. Assuming that there is no slipping between the sphere and the surface, draw the necessary free-body diagram and derive the differential equation of motion.
16. Consider the system shown in Figure 5.74. A uniform solid cylinder of mass m , radius R , and length L is fitted with a frictionless axle along the cylinder's long axis. A spring of stiffness k is attached to a bracket connected to the axle. Assume that the cylinder rolls without slipping on a horizontal surface. Draw the necessary free-body diagram and derive the differential equation of motion.

**FIGURE 5.73**

Problem 15.

**FIGURE 5.74**

Problem 16.

5.4 Mixed Systems: Translational and Rotational

For a system involving both translational and rotational motions, Newton's second law and the moment equation can be used to obtain the model. This section provides only one method to obtain the differential equation of motion by using the force and moment equations, which may be applied in different ways.

5.4.1 Force and Moment Equations

Consider a mechanical system in plane motion, which involves translations along the x and y directions and rotation about one axis perpendicular to the x - y plane. For a system of a single mass, applying Newton's second law in the translational directions gives the force equations

$$\sum F_x = ma_{Cx}, \quad \sum F_y = ma_{Cy} \quad (5.39)$$

The moment equation is in the form of

$$\sum M_C = I_C \alpha \quad (5.40)$$

or

$$\sum M_P = I_C \alpha + M_{\text{eff_}ma_C} \quad (5.41)$$

where the symbol $M_{\text{eff_}ma_C}$ is used to represent the effective moment caused by the fictitious force ma_C . Although Equation 5.41 looks different from the general moment Equation 5.31 given in Section 5.3, they are essentially equivalent. Figure 5.75 shows a rigid body in plane motion. Applying Equation 5.31 gives the net moment about point P as $\sum \mathbf{M}_P = I_P \alpha + m \mathbf{r}_{C/P} \times \mathbf{a}_P$. According to the parallel-axis theorem, we have $I_P = I_C + m r_{C/P}^2$. Then, $I_P \alpha = I_C \alpha + m r_{C/P}^2 \alpha$, and it follows the direction of α , that is, the counterclockwise direction. Also, from the kinematics of rigid bodies, the acceleration of point P is $\mathbf{a}_P = \mathbf{a}_C + \mathbf{a}_{P/C,t} + \mathbf{a}_{P/C,n}$ as shown in Figure 5.76. Note that $m \mathbf{r}_{C/P} \times \mathbf{a}_C = \mathbf{r}_{C/P} \times (m \mathbf{a}_C)$, which can be considered as the effective moment caused by the fictitious force $m \mathbf{a}_C$ and is denoted as $M_{\text{eff_}ma_C}$. The tangential component of the relative acceleration $\mathbf{a}_{P/C,t}$ is perpendicular to the position vector $\mathbf{r}_{C/P}$ with a magnitude of $\alpha r_{C/P}$. Thus, the magnitude of the cross product $m \mathbf{r}_{C/P} \times \mathbf{a}_{P/C,t}$ is $m r_{C/P}^2 \alpha$ and

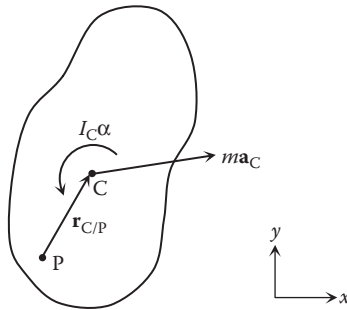
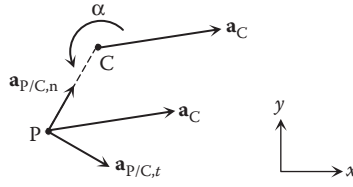


FIGURE 5.75
A rigid body in plane motion.

**FIGURE 5.76**

A kinematic diagram of $\mathbf{a}_P$ and $\mathbf{a}_C$.

its direction is clockwise. The normal component $\mathbf{a}_{P/C,n}$ is parallel to the position vector $\mathbf{r}_{C/P}$, and thus, $m\mathbf{r}_{C/P} \times \mathbf{a}_{P/C,n} = 0$. Assuming that the positive direction is counterclockwise, we have $\Sigma M_P = I_C \alpha + m r_{C/P}^2 \alpha + M_{\text{eff_mac}} - m r_{C/P}^2 \alpha$, which is the same as Equation 5.41.

For a system of multiple masses, the force equations become

$$\sum F_x = \sum_{i=1}^n m_i (a_{Ci})_x, \quad \sum F_y = \sum_{i=1}^n m_i (a_{Ci})_y \quad (5.42)$$

where n is the number of masses and a_{Ci} is the acceleration of the mass center of the i th mass. Because it may be a challenge to find the mass center for the entire system of multiple masses, we use an arbitrary point P to sum the moments,

$$\sum M_P = \sum_{i=1}^n I_{Ci} \alpha_i + \sum_{i=1}^n M_{\text{eff_}m_i a_{Ci}} \quad (5.43)$$

where:

I_{Ci} is the moment of inertia of the i th mass about its center of mass

α_i is the angular acceleration of the i th mass

To derive the differential equation of motion correctly, we recommend drawing two diagrams before applying the force and moment equations. One is the free-body diagram that shows all external forces and moments applied to the system, and the other is the kinematic diagram that indicates the acceleration at the mass center of each mass. The left-hand sides of the force and moment Equations 5.42 and 5.43 are written based on the free-body diagram, and the right-hand sides are written based on the kinematic diagram.

Example 5.13: A Lever Mechanism

Consider the system shown in Figure 5.77, in which a lever arm has a force applied on one side and a spring–damper combination on the other side with a suspended mass. When $\theta = 0$ and $\dot{f} = 0$, the system is at static equilibrium. Assume that the lever arm can be approximated as a uniform slender rod. Draw the free-body diagram of the lever arm and the suspended mass. Derive the differential equations of motion for small angles θ .

Solution

This is a mixed system with two masses. The suspended block undergoes translational motion along the vertical direction, and the motion of the rod is pure rotation about

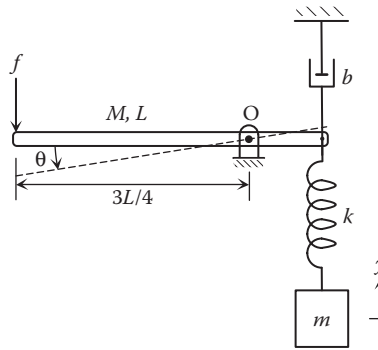


FIGURE 5.77
A lever mechanism.

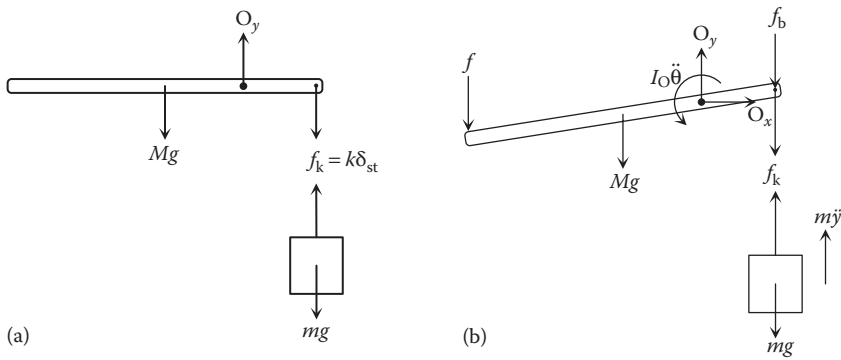


FIGURE 5.78
A lever mechanism: (a) free-body diagram at static equilibrium and (b) free-body diagram and kinematic diagram at dynamic position.

point O that is fixed. We choose the displacement of the block y and the angular displacement of the rod θ as the generalized coordinates. Note that y is measured from the static equilibrium position, which is set as the origin. The free-body diagrams and the kinematic diagram of the system are shown in [Figure 5.78](#).

At static equilibrium, for the block, we have

$$\begin{aligned} + \uparrow: \quad \sum F_y &= 0 \\ k\delta_{st} - mg &= 0 \\ k\delta_{st} &= mg \end{aligned}$$

and for the arm,

$$\begin{aligned} + \curvearrowright: \quad \sum M_O &= 0 \\ Mg \cdot \frac{L}{4} - k\delta_{st} \cdot \frac{L}{4} &= 0 \\ Mg &= k\delta_{st} \end{aligned}$$

where δ_{st} is the static deformation of the spring.

When a force f is applied on one side of the rod, the deformation of the spring caused by the rotation of the rod can be approximated as $L\theta/4$ for small angles θ . Assuming that the block and the rod are displaced in their positive directions and $L\theta/4 > y > 0$, the spring is in tension and the magnitude of the spring force is $f_k = k(L\theta/4 - y + \delta_{st})$. The magnitude of the damping force is $f_b = bL\dot{\theta}/4$. The free body diagram and the kinematic diagram of the system are shown in Figure 5.78b. Note that the acceleration components $m\ddot{y}$ and $I_O\ddot{\theta}$ are shown together with the forces in the free-body diagram. For complex mechanical systems, we recommend to draw a separate kinematic diagram.

For the block (translation only), applying the force equation in the y direction gives

$$+\uparrow y: \sum F_y = ma_{Cy}$$

$$k\left(\frac{L}{4}\theta - y + \delta_{st}\right) - mg = m\ddot{y}$$

For the rod (rotation only), applying the moment equation about the fixed point O gives

$$+\curvearrowright: \sum M_O = I_O\alpha$$

$$f \cdot \frac{3L}{4} \cos\theta + Mg \cdot \frac{L}{4} \cos\theta - k\left(\frac{L}{4}\theta - y + \delta_{st}\right) \cdot \frac{L}{4} \cos\theta - b\frac{L}{4}\dot{\theta} \cdot \frac{L}{4} \cos\theta = I_O\ddot{\theta}$$

where I_O can be obtained using parallel-axis theorem,

$$I_O = \frac{1}{12}ML^2 + M\left(\frac{L}{4}\right)^2 = \frac{7}{48}ML^2$$

For small angular motions, $\cos\theta \approx 1$. Introducing the static equilibrium conditions and rearranging the two equations give

$$m\ddot{y} + ky - \frac{kL}{4}\theta = 0$$

$$\frac{7}{48}ML^2\ddot{\theta} + \frac{bL^2}{16}\dot{\theta} + \frac{kL^2}{16}\theta - \frac{kL}{4}y = \frac{3fL}{4}$$

Example 5.14: A Cart–Inverted-Pendulum System

Consider the mechanical system shown in Figure 5.79, where a uniform rod of mass M and length L is pivoted on a cart of mass m . An external force f is applied to the cart. Assume that the pendulum is constrained to move in a vertical plane, and the cart moves without slipping along a horizontal line. Denote the displacement of the cart as x and the angular displacement of the pendulum as θ .

- Draw the necessary free-body diagram and kinematic diagram, and derive the nonlinear equations of motion.
- Linearize the equations of motion for small angular motions, and determine the state-space form with x and θ as the outputs.

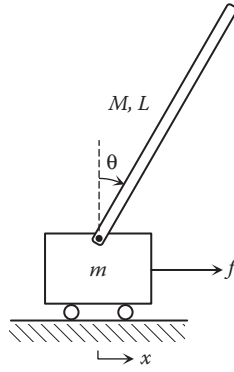


FIGURE 5.79
A cart-inverted-pendulum system.

Solution

- a. This is a mixed system with two masses. The motion of the cart is purely translational, and the inverted pendulum undergoes both translation and rotation. The free-body diagram and the kinematic diagram are given in Figure 5.80. Note that the acceleration of the mass center of the rod consists of three components. On the one hand, the pendulum moves together with the cart along the horizontal line at an acceleration of $\ddot{x}$. On the other hand, it rotates about the pivot P on the cart. The relative rotational accelerations include the tangential component $\frac{L}{2}\ddot{\theta}$ and the normal component $\frac{L}{2}\dot{\theta}^2$.

Applying the force equation to the whole system along the x direction gives

$$+\rightarrow x: \sum F_x = \sum_{i=1}^2 m_i (a_{Ci})_x$$

$$f = m\ddot{x} + M\ddot{x} - M\frac{L}{2}\dot{\theta}^2 \cdot \sin\theta + M\frac{L}{2}\ddot{\theta} \cdot \cos\theta$$

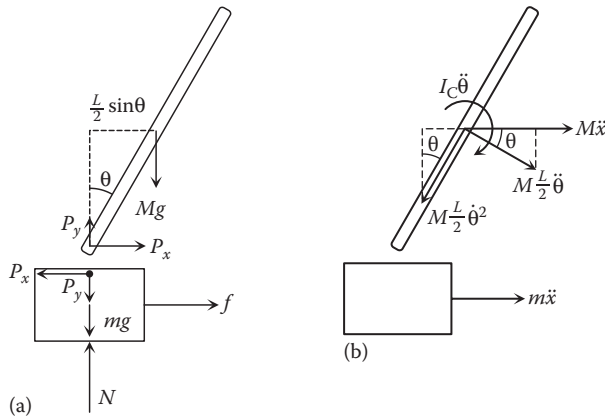


FIGURE 5.80
A cart-inverted-pendulum system: (a) free-body diagram and (b) kinematic diagram.

The forces at the pivot are canceled out because they are internal forces between the cart and the pendulum. Applying the moment equation to the pendulum about point P results in

$$+ \curvearrowright: \sum M_P = I_C \alpha + M_{\text{eff_mac}}$$

$$Mg \cdot \frac{L}{2} \sin \theta = \frac{1}{12} ML^2 \ddot{\theta} + M \ddot{x} \cdot \frac{L}{2} \cos \theta + M \frac{L}{2} \ddot{\theta} \cdot \frac{L}{2}$$

Rearranging the two equations into the standard input–output form

$$(m + M) \ddot{x} + \frac{1}{2} ML \ddot{\theta} \cos \theta - \frac{1}{2} ML \dot{\theta}^2 \sin \theta = f$$

$$\frac{1}{3} ML^2 \ddot{\theta} + \frac{1}{2} ML \ddot{x} \cos \theta - \frac{1}{2} MgL \sin \theta = 0$$

- b. For small angular motions, $\cos \theta \approx 1$, $\sin \theta \approx \theta$, $\dot{\theta}^2 \theta \approx 0$ (see [Section 4.6](#)). The linearized equations are

$$(m + M) \ddot{x} + \frac{1}{2} ML \ddot{\theta} = f$$

$$\frac{1}{3} ML^2 \ddot{\theta} + \frac{1}{2} ML \ddot{x} - \frac{1}{2} MgL \theta = 0$$

The state, the input, and the output are specified as

$$\mathbf{x} = \begin{Bmatrix} x_1 \\ x_2 \\ x_3 \\ x_4 \end{Bmatrix} = \begin{Bmatrix} x \\ \theta \\ \dot{x} \\ \dot{\theta} \end{Bmatrix}, \quad u = f, \quad \mathbf{y} = \begin{Bmatrix} x \\ \theta \end{Bmatrix}$$

We then take the time derivative of each state variable. For the first two,

$$\dot{x}_1 = \dot{x} = x_3$$

$$\dot{x}_2 = \dot{\theta} = x_4$$

Note that the two linearized equations are coupled. The derivatives $\ddot{x}$ and $\ddot{\theta}$ cannot be solved using only one of the equations. To find $\dot{x}_3$ and $\dot{x}_4$, we rewrite the two linearized differential equations as

$$\begin{bmatrix} m + M & \frac{1}{2} ML \\ \frac{1}{2} ML & \frac{1}{3} ML^2 \end{bmatrix} \begin{Bmatrix} \ddot{x} \\ \ddot{\theta} \end{Bmatrix} = \begin{Bmatrix} f \\ \frac{1}{2} MgL \theta \end{Bmatrix}$$

from which $\ddot{x}$ and $\ddot{\theta}$ can be solved using Cramer's rule:

$$\ddot{x} = \frac{(1/3)ML^2 f - (1/4)M^2 L^2 g \theta}{(1/12)ML^2 (M + 4m)}$$

$$\ddot{\theta} = \frac{-(1/2)ML f + (M + m)(1/2)MgL \theta}{(1/12)ML^2 (M + 4m)}$$

Simplifying the equations gives

$$\dot{x}_3 = \frac{4}{M+4m}u - \frac{3Mg}{M+4m}x_2$$

$$\dot{x}_4 = \frac{-6}{L(M+4m)}u + \frac{6(M+m)g}{L(M+4m)}x_2$$

Thus, the state-space equation in matrix form is

$$\begin{Bmatrix} \dot{x}_1 \\ \dot{x}_2 \\ \dot{x}_3 \\ \dot{x}_4 \end{Bmatrix} = \begin{bmatrix} 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 1 \\ 0 & -\frac{3Mg}{M+4m} & 0 & 0 \\ 0 & \frac{6(M+m)g}{L(M+4m)} & 0 & 0 \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \\ x_3 \\ x_4 \end{Bmatrix} + \begin{bmatrix} 0 \\ 0 \\ \frac{4}{M+4m} \\ -\frac{6}{L(M+4m)} \end{bmatrix} u$$

$$\mathbf{y} = \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \\ x_3 \\ x_4 \end{Bmatrix} + \begin{bmatrix} 0 \\ 0 \end{bmatrix} u$$

Note that in Part (a), Equation 5.31 instead of 5.41 can also be applied to the pendulum to derive the second equation of motion. The related free-body diagram and the kinematic diagram are shown in [Figure 5.81](#). Summing all externally applied moments about the pivot P, which moves together with the cart with an acceleration of $\ddot{x}$, gives

$$+ \curvearrowright: \sum \mathbf{M}_P = I_P \alpha + m \mathbf{r}_{C/P} \times \mathbf{a}_P$$

$$Mg \cdot \frac{L}{2} \sin \theta = \frac{1}{3} ML^2 \ddot{\theta} + M \frac{L}{2} \ddot{x} \sin \left(\frac{\pi}{2} - \theta \right)$$

$$\frac{1}{3} ML^2 \ddot{\theta} + \frac{1}{2} ML \ddot{x} \cos \theta - \frac{1}{2} MgL \sin \theta = 0$$

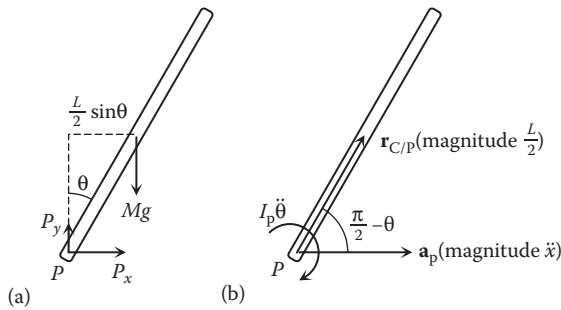


FIGURE 5.81

Applying Equation 5.31 to the pendulum: (a) free-body diagram and (b) kinematic diagram.

5.4.2 Energy Method

The differential equation of motion of a mechanical system can be obtained using the force/moment approach, which is based on Newtonian mechanics. Free-body diagrams are necessary in order to apply the force and moment equations correctly. An alternative way of obtaining the differential equation of motion is to use the energy method based on analytical mechanics.

For a mass–spring system with negligible friction and damping, the principle of conservation of energy states that

$$T + V = \text{constant} \quad (5.44)$$

or

$$\frac{d}{dt}(T + V) = 0 \quad (5.45)$$

where:

T is the kinetic energy

V is the potential energy, which includes the gravitational potential energy and the elastic potential energy

The expression for the kinetic energy of a translational or rotational mass element was given in [Section 5.1](#). In general, the kinetic energy of a rigid body in plane motion is

$$T = \frac{1}{2}mv_C^2 + \frac{1}{2}I_C\omega^2 \quad (5.46)$$

where:

v_C is the velocity of the mass center C of the body

ω is the angular velocity of the body

Note that the kinetic energy of a rigid body in plane motion can be separated into two parts: (1) the kinetic energy associated with the translational motion of the mass center C of the body, and (2) the kinetic energy associated with the rotation of the body about the mass center C . If a rigid body rotates about a fixed point O with an angular velocity ω , the kinetic energy reduces to $T = (1/2)I_O\omega^2$.

Example 5.15: A Pulley System

Consider the pulley system shown in [Figure 5.82](#). A block of mass m is connected to a translational spring of stiffness k through a cable, which passes by a pulley with mass m_p and radius r . The pulley rotates about the fixed mass center O . The moment of inertia of the pulley about its mass center is I_O . Use the energy method to derive the equation of motion.

Solution

The system has two mass elements, one translational block of mass and one pulley rotating about its fixed mass center. Their motions are related to each other by the geometric constraint, $x = r\theta$. The kinetic energy of the system is

$$T = T_{\text{block}} + T_{\text{pulley}} = \frac{1}{2}m\dot{x}^2 + \frac{1}{2}I_O\dot{\theta}^2$$

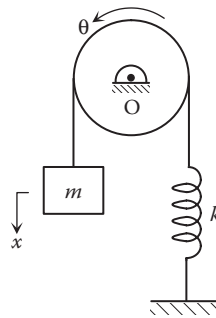


FIGURE 5.82
A pulley system.

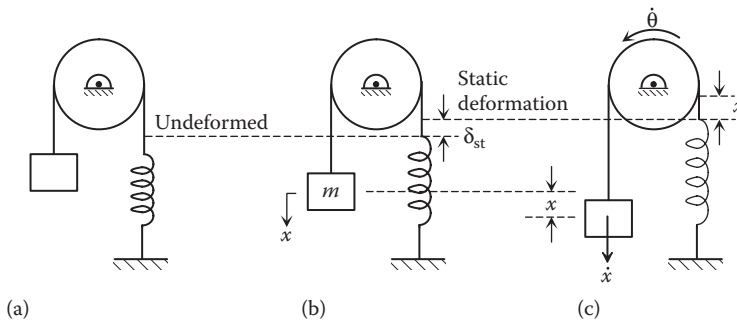


FIGURE 5.83
A pulley system: (a) undeformed position, (b) static equilibrium position, and (c) dynamic position.

The undeformed position, static equilibrium position, and dynamic position of the mass–spring system are shown in [Figure 5.83](#).

Choosing the static equilibrium position as the datum for the gravitational potential energy, we have

$$V_g = -mgx$$

The elastic potential energy is

$$V_e = \frac{1}{2}k(x + \delta_{st})^2$$

The total potential energy is

$$V = V_g + V_e = -mgx + \frac{1}{2}kx^2 + \frac{1}{2}k\delta_{st}^2 + kx\delta_{st}$$

Because of the static equilibrium condition, $mg = k\delta_{st}$, the expression of potential energy becomes

$$V = \frac{1}{2}kx^2 + \frac{1}{2}k\delta_{st}^2$$

The total energy of the system is

$$T + V = \frac{1}{2}m\dot{x}^2 + \frac{1}{2}I_O\dot{\theta}^2 + \frac{1}{2}kx^2 + \frac{1}{2}k\delta_{st}^2$$

for which the time derivative is

$$\frac{d}{dt}(T + V) = m\dot{x}\ddot{x} + I_O\dot{\theta}\ddot{\theta} + kx\dot{x}$$

Using the geometric constraint, $x = r\theta$, which implies that $\dot{x} = r\dot{\theta}$ and $\ddot{x} = r\ddot{\theta}$, we can rewrite the above-mentioned equation as

$$\frac{d}{dt}(T + V) = mr^2\dot{\theta}\ddot{\theta} + I_O\dot{\theta}\ddot{\theta} + kr^2\dot{\theta}$$

Applying the principle of conservation of energy, we find

$$(I_O + mr^2)\ddot{\theta} + kr^2\dot{\theta} = 0$$

Note that the mechanical system in Example 5.15 is a single-degree-of-freedom system, which requires only one generalized coordinate (e.g., x or θ) to describe the system dynamics. The example shows that if we can obtain the expression of $T + V$ in terms of the generalized coordinate, the equation of motion can be derived by taking the time derivative and setting it equal to zero.

For an n -degree-of-freedom system, n independent equations of motion can be derived using Lagrange's formulation, which is applicable to both conservative and nonconservative systems. One of the forms of Lagrange's equations for a conservative system is

$$\frac{d}{dt}\left(\frac{\partial T}{\partial \dot{q}_i}\right) - \frac{\partial T}{\partial q_i} + \frac{\partial V}{\partial q_i} = 0, \quad i = 1, 2, \dots, n \quad (5.47)$$

where:

q_i is the i th generalized coordinate

n is the total number of independent generalized coordinates

In general, the kinetic energy is a function of the generalized displacements and the generalized velocities,

$$T = T(q_1, q_2, \dots, q_n, \dot{q}_1, \dot{q}_2, \dots, \dot{q}_n) \quad (5.48)$$

The potential energy is a function of the generalized displacements,

$$V = V(q_1, q_2, \dots, q_n) \quad (5.49)$$

Example 5.16: A Double Pendulum

Consider the double pendulum in [Figure 5.84](#), where two point masses of equal mass m are attached to two rigid links of equal length L . The links are assumed to be massless. The motion of the system is constrained in the vertical plane. Neglecting friction, derive the equations of motion by using Lagrange's equations. Assume small angles for θ_1 and θ_2 .

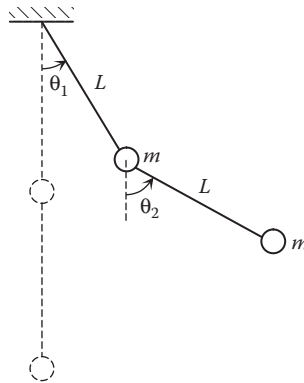


FIGURE 5.84
A double-pendulum system.

Solution

The system is only subjected to gravitational forces, and it is a conservative system. The dynamics of the system can be described using two independent angular displacement coordinates, θ_1 and θ_2 . The kinetic energy of the system is

$$T = \frac{1}{2}mv_1^2 + \frac{1}{2}mv_2^2$$

From kinematics,

$$v_1 = L\dot{\theta}_1$$

and, as shown in [Figure 5.85a](#),

$$v_2^2 = (L\dot{\theta}_1)^2 + (L\dot{\theta}_2)^2 + 2L^2\dot{\theta}_1\dot{\theta}_2 \cos(\theta_2 - \theta_1)$$

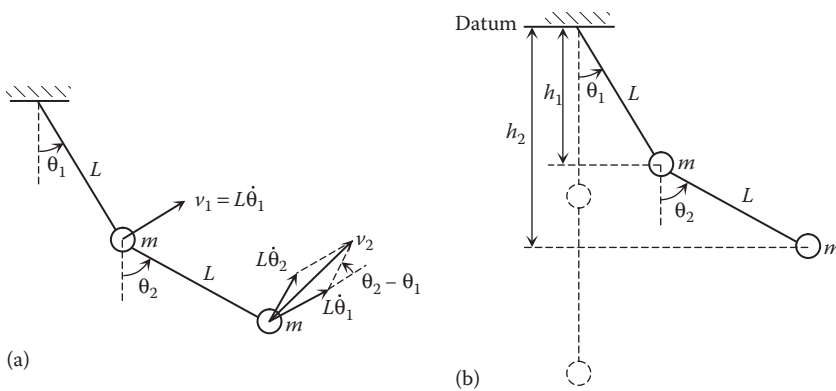


FIGURE 5.85
A double-pendulum: (a) kinematic diagram and (b) positions of the mass centers.

which is obtained by applying the law of cosines. Thus,

$$T = \frac{1}{2}m(\dot{L}\dot{\theta}_1)^2 + \frac{1}{2}m\left[(\dot{L}\dot{\theta}_1)^2 + (\dot{L}\dot{\theta}_2)^2 + 2L^2\dot{\theta}_1\dot{\theta}_2\cos(\theta_2 - \theta_1)\right]$$

Note that no spring elements are involved in the system. This implies that $V_e = 0$ and $V = V_g$. Using the datum defined in [Figure 5.85b](#), we can obtain the gravitational potential energy

$$V_g = -mgh_1 - mgh_2$$

where

$$h_1 = L\cos\theta_1$$

$$h_2 = L\cos\theta_1 + L\cos\theta_2$$

Thus,

$$V = -2mgL\cos\theta_1 - mgL\cos\theta_2$$

We then apply Lagrange's equations

$$\frac{d}{dt}\left(\frac{\partial T}{\partial \dot{q}_i}\right) - \frac{\partial T}{\partial q_i} + \frac{\partial V}{\partial q_i} = 0, \quad i = 1, 2$$

For $i = 1$, $q_1 = \theta_1$, $\dot{q}_1 = \dot{\theta}_1$,

$$\frac{\partial T}{\partial \dot{\theta}_1} = 2mL^2\dot{\theta}_1 + mL^2\dot{\theta}_2\cos(\theta_2 - \theta_1)$$

$$\frac{d}{dt}\left(\frac{\partial T}{\partial \dot{\theta}_1}\right) = 2mL^2\ddot{\theta}_1 + mL^2\ddot{\theta}_2\cos(\theta_2 - \theta_1) - mL^2\dot{\theta}_2\sin(\theta_2 - \theta_1)(\dot{\theta}_2 - \dot{\theta}_1)$$

$$\frac{\partial T}{\partial \theta_1} = -mL^2\dot{\theta}_1\dot{\theta}_2\sin(\theta_2 - \theta_1)(-1)$$

$$\frac{\partial V}{\partial \theta_1} = 2mgL\sin\theta_1$$

Substituting into Lagrange's equation results in

$$\frac{d}{dt}\left(\frac{\partial T}{\partial \dot{\theta}_1}\right) - \frac{\partial T}{\partial \theta_1} + \frac{\partial V}{\partial \theta_1} = 2mL^2\ddot{\theta}_1 + mL^2\ddot{\theta}_2\cos(\theta_2 - \theta_1) - mL^2\dot{\theta}_2^2\sin(\theta_2 - \theta_1) + 2mgL\sin\theta_1 = 0$$

Similarly, for $i = 2$, $q_2 = \theta_2$, $\dot{q}_2 = \dot{\theta}_2$,

$$\begin{aligned}
\frac{\partial T}{\partial \dot{\theta}_2} &= mL^2 \dot{\theta}_2 + mL^2 \dot{\theta}_1 \cos(\theta_2 - \theta_1) \\
\frac{d}{dt} \left(\frac{\partial T}{\partial \dot{\theta}_2} \right) &= mL^2 \ddot{\theta}_2 + mL^2 \ddot{\theta}_1 \cos(\theta_2 - \theta_1) - mL^2 \dot{\theta}_1 \sin(\theta_2 - \theta_1) (\dot{\theta}_2 - \dot{\theta}_1) \\
\frac{\partial T}{\partial \theta_2} &= -mL^2 \dot{\theta}_1 \dot{\theta}_2 \sin(\theta_2 - \theta_1) \\
\frac{\partial V}{\partial \theta_2} &= mgL \sin \theta_2
\end{aligned}$$

Substituting into Lagrange's equation gives

$$\frac{d}{dt} \left(\frac{\partial T}{\partial \dot{\theta}_2} \right) - \frac{\partial T}{\partial \theta_2} + \frac{\partial V}{\partial \theta_2} = mL^2 \ddot{\theta}_2 + mL^2 \ddot{\theta}_1 \cos(\theta_2 - \theta_1) + mL^2 \dot{\theta}_1^2 \sin(\theta_2 - \theta_1) + mgL \sin \theta_2 = 0$$

For small motions ($\theta_1 \approx 0$ and $\theta_2 \approx 0$), the two differential equations of motion are linearized as

$$\begin{aligned}
2mL^2 \ddot{\theta}_1 + mL^2 \ddot{\theta}_2 + 2mgL \theta_1 &= 0 \\
mL^2 \ddot{\theta}_1 + mL^2 \ddot{\theta}_2 + mgL \theta_2 &= 0
\end{aligned}$$

or in second-order matrix form

$$\begin{bmatrix} 2mL^2 & mL^2 \\ mL^2 & mL^2 \end{bmatrix} \begin{Bmatrix} \ddot{\theta}_1 \\ \ddot{\theta}_2 \end{Bmatrix} + \begin{bmatrix} 2mgL & 0 \\ 0 & mgL \end{bmatrix} \begin{Bmatrix} \theta_1 \\ \theta_2 \end{Bmatrix} = \begin{Bmatrix} 0 \\ 0 \end{Bmatrix}$$

For a nonconservative system, the form of Lagrange's equations given in Equation 5.47 is changed to

$$\frac{d}{dt} \left(\frac{\partial T}{\partial \dot{q}_i} \right) - \frac{\partial T}{\partial q_i} + \frac{\partial V}{\partial q_i} = Q_i, \quad i = 1, 2, \dots, n \quad (5.50)$$

where Q_i 's represent nonconservative forces or moments associated with generalized coordinates, known as generalized forces. Typical examples of nonconservative forces include the damping force, the friction force, and other external forces by which the work done is not zero. Note that both the gravitational force and the spring force are conservative forces, which are taken into account when finding the expression of the potential energy V .

The generalized force associated with the i th generalized coordinate can be determined by

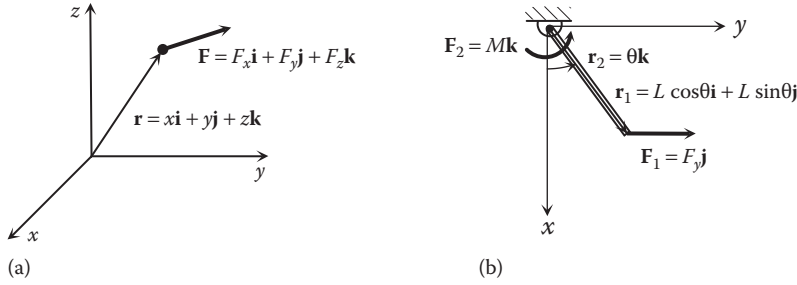
$$Q_i = \sum_{j=1}^N \mathbf{F}_j \frac{\partial \mathbf{r}_j}{\partial q_i} \quad j = 1, 2, \dots, N \quad (5.51)$$

where:

$\mathbf{F}_j$ is the nonconservative force vector applied to the system at the j th point

$\mathbf{r}_j$ is the location vector of the j th point with respect to a fixed coordinate system

N is the number of nonconservative forces applied to the system

**FIGURE 5.86**

Example of generalized forces: (a) translational motion and (b) rotational motion.

For instance, the point mass in [Figure 5.86a](#) is subjected to a force $\mathbf{F}$. It moves in the three-dimensional space, and $\mathbf{r}$ is the position vector. Choosing x , y , and z as the generalized coordinates, we can find the associated generalized forces by applying Equation 5.51,

$$Q_1 = \mathbf{F} \frac{\partial \mathbf{r}}{\partial q_1} = (F_x \mathbf{i} + F_y \mathbf{j} + F_z \mathbf{k}) \frac{\partial}{\partial x} (x\mathbf{i} + y\mathbf{j} + z\mathbf{k}) = (F_x \mathbf{i} + F_y \mathbf{j} + F_z \mathbf{k}) \mathbf{i} = F_x$$

$$Q_2 = \mathbf{F} \frac{\partial \mathbf{r}}{\partial q_2} = (F_x \mathbf{i} + F_y \mathbf{j} + F_z \mathbf{k}) \frac{\partial}{\partial y} (x\mathbf{i} + y\mathbf{j} + z\mathbf{k}) = (F_x \mathbf{i} + F_y \mathbf{j} + F_z \mathbf{k}) \mathbf{j} = F_y$$

$$Q_3 = \mathbf{F} \frac{\partial \mathbf{r}}{\partial q_3} = (F_x \mathbf{i} + F_y \mathbf{j} + F_z \mathbf{k}) \frac{\partial}{\partial z} (x\mathbf{i} + y\mathbf{j} + z\mathbf{k}) = (F_x \mathbf{i} + F_y \mathbf{j} + F_z \mathbf{k}) \mathbf{k} = F_z$$

The example in [Figure 5.86b](#) shows how to find the generalized force if the rotational motion is involved. A pendulum of length L is subjected to a horizontal force and a torque at the pivot. Within the chosen coordinate system x - y - z , the torque can be expressed as $M\mathbf{k}$ and the resulting position vector at the pivot can be expressed as $\theta\mathbf{k}$. With the angular displacement θ as the generalized coordinate, applying Equation 5.51 gives

$$\begin{aligned} Q &= \sum_{j=1}^2 \mathbf{F}_j \frac{\partial \mathbf{r}_j}{\partial q} = (F_j) \frac{\partial}{\partial \theta} (L \cos \theta \mathbf{i} + L \sin \theta \mathbf{j}) + (M\mathbf{k}) \frac{\partial}{\partial \theta} (\theta \mathbf{k}) \\ &= (F_j) (-L \sin \theta \mathbf{i} + L \cos \theta \mathbf{j}) + (M\mathbf{k}) \mathbf{k} = FL \cos \theta + M \end{aligned}$$

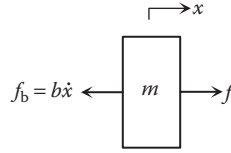
Note that Equation 5.51 is derived based on the principle of virtual work, which is beyond the scope of this text. The interested reader can refer to [13] for more details.

Example 5.17: A Single-Degree-of-Freedom Mass–Spring–Damper System (Energy Method)

Reconsider the simple mass–spring–damper system, as shown in [Figure 5.29a](#). Derive the equation of motion by using Lagrange's equations.

Solution

The displacement x is chosen as the generalized coordinate, and the origin is set at the static equilibrium position. Other than the spring force, the system is subjected to the damping force f_b and the external force f (as shown in [Figure 5.87](#)), and it is a nonconservative system.

**FIGURE 5.87**

The nonconservative forces applied to a mass–spring–damper system.

The kinetic energy and the potential energy of the system are $T = \frac{1}{2}m\dot{x}^2$ and $V = \frac{1}{2}kx^2$, respectively. The generalized force is

$$Q = \sum_{j=1}^2 \mathbf{F}_j \frac{\partial \mathbf{r}_j}{\partial \mathbf{x}} = (\mathbf{f}\mathbf{i}) \frac{\partial}{\partial x}(x\mathbf{i}) + (-b\dot{x}\mathbf{i}) \frac{\partial}{\partial x}(x\mathbf{i}) = f - b\dot{x}$$

Applying the Lagrange's equation

$$\frac{d}{dt} \left(\frac{\partial T}{\partial \dot{x}} \right) - \frac{\partial T}{\partial x} + \frac{\partial V}{\partial x} = Q$$

gives the equation of motion of the system

$$m\ddot{x} + kx = f - b\dot{x}$$

or

$$m\ddot{x} + b\dot{x} + kx = f$$

Example 5.18: A Cart–Inverted-Pendulum System (Energy Method)

Reconsider the mechanical system, as shown in [Figure 5.79](#). Derive the equations of motion of the system by using Lagrange's equations.

Solution

As shown in [Figure 5.88a](#), the motion of the cart is purely translational and

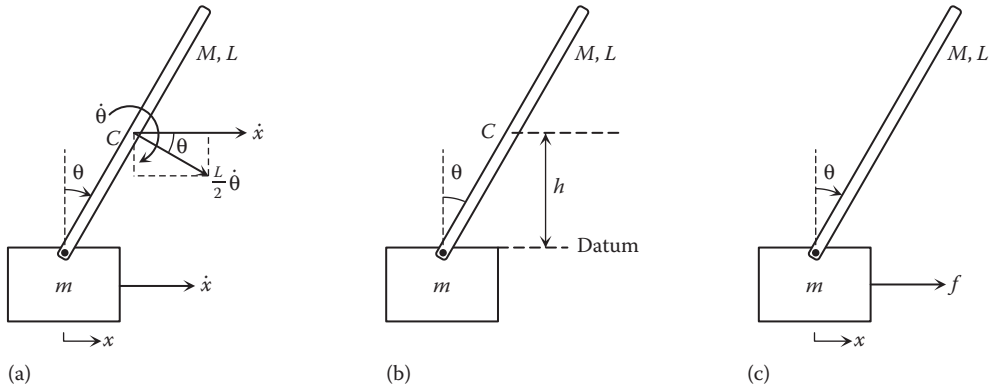
$$T_{\text{cart}} = \frac{1}{2}m\dot{x}^2$$

The inverted pendulum undergoes both translation and rotation and

$$T_{\text{pendulum}} = \frac{1}{2}Mv_C^2 + \frac{1}{2}I_C\dot{\theta}^2$$

where the velocity at the center of mass can be found by

$$v_C^2 = \left(\dot{x} + \frac{L}{2}\dot{\theta}\cos\theta \right)^2 + \left(\frac{L}{2}\dot{\theta}\sin\theta \right)^2 = \dot{x}^2 + L\dot{x}\dot{\theta}\cos\theta + \frac{1}{4}L^2\dot{\theta}^2$$

**FIGURE 5.88**

A cart–inverted–pendulum system: (a) kinematic diagram, (b) position of the mass center of the pendulum, and (c) nonconservative force.

and the mass moment of inertia $I_C = \frac{1}{12} ML^2$. Thus, the kinetic energy of the system is

$$T = T_{\text{cart}} + T_{\text{pendulum}} = \frac{1}{2}(m + M)\dot{x}^2 + \frac{1}{6}ML^2\dot{\theta}^2 + \frac{1}{2}ML\dot{x}\dot{\theta}\cos\theta$$

Note that no spring elements are involved in the system. This implies that $V_e = 0$ and $V = V_g$. Using the datum defined in Figure 5.88b, we can obtain the gravitational potential energy

$$V_g = Mgh = Mg \frac{L}{2} \cos\theta$$

As shown in Figure 5.88c, the system is subjected to a horizontal force f , by which the work done is not zero. Thus, it is a nonconservative system. Same as Example 5.14, the displacement of the cart, x , and the angular displacement of the pendulum, θ , are chosen as the generalized coordinates. We then apply Lagrange's equations

$$\frac{d}{dt} \left(\frac{\partial T}{\partial \dot{q}_i} \right) - \frac{\partial T}{\partial q_i} + \frac{\partial V}{\partial q_i} = Q_i, \quad i = 1, 2$$

For $i = 1$, $q_1 = x$, $\dot{q}_1 = \dot{x}$,

$$\begin{aligned} \frac{\partial T}{\partial \dot{x}} &= (m + M)\dot{x} + \frac{1}{2}ML\dot{\theta}\cos\theta \\ \frac{d}{dt} \left(\frac{\partial T}{\partial \dot{x}} \right) &= (m + M)\ddot{x} + \frac{1}{2}ML\ddot{\theta}\cos\theta - \frac{1}{2}ML\dot{\theta}\sin\theta\dot{\theta} \\ \frac{\partial T}{\partial x} &= 0 \\ \frac{\partial V}{\partial x} &= 0 \end{aligned}$$

Applying Equation 5.51 gives the generalized force

$$Q_1 = \mathbf{F} \frac{\partial \mathbf{r}}{\partial x} = (\mathbf{f}) \frac{\partial}{\partial x} (x\mathbf{i}) = f$$

Substituting into Lagrange's equation results in

$$(m + M)\ddot{x} + \frac{1}{2}ML\ddot{\theta}\cos\theta - \frac{1}{2}ML\dot{\theta}^2\sin\theta = f$$

Similarly, for $i = 2, q_2 = \theta, \dot{q}_2 = \dot{\theta}$,

$$\begin{aligned}\frac{\partial T}{\partial \dot{\theta}} &= \frac{1}{3}ML^2\dot{\theta} + \frac{1}{2}ML\dot{x}\cos\theta \\ \frac{d}{dt}\left(\frac{\partial T}{\partial \dot{\theta}}\right) &= \frac{1}{3}ML^2\ddot{\theta} + \frac{1}{2}ML\ddot{x}\cos\theta - \frac{1}{2}ML\dot{x}\sin\theta\dot{\theta} \\ \frac{\partial T}{\partial \theta} &= -\frac{1}{2}ML\dot{x}\dot{\theta}\sin\theta \\ \frac{\partial V}{\partial \theta} &= -\frac{1}{2}MgL\sin\theta\end{aligned}$$

Applying Equation 5.51 gives the generalized force

$$Q_2 = \mathbf{F} \frac{\partial \mathbf{r}}{\partial \theta} = (\mathbf{f}\mathbf{i}) \frac{\partial}{\partial \theta}(x\mathbf{i}) = 0$$

Note that the external force f does not generate a moment on the pendulum. Substituting into Lagrange's equation gives

$$\frac{1}{3}ML^2\ddot{\theta} + \frac{1}{2}ML\ddot{x}\cos\theta - \frac{1}{2}MgL\sin\theta = 0$$

The result is the same as the one obtained previously in Example 5.14.

PROBLEM SET 5.4

1. For the pulley system in Example 5.15, draw the free-body diagram and kinematic diagram, and derive the equation of motion by using the force/moment approach.
2. The double pulley system shown in Figure 5.89 has an inner radius of r_1 and an outer radius of r_2 . The mass moment of inertia of the pulley about point O is I_O . A translational spring of stiffness k and a block of mass m are suspended by cables wrapped around the pulley, as shown. Draw the free-body diagram and kinematic diagram, and derive the equation of motion by using the force/moment approach.

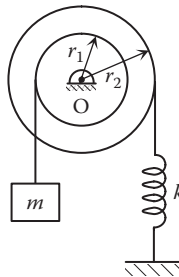


FIGURE 5.89
Problem 2.

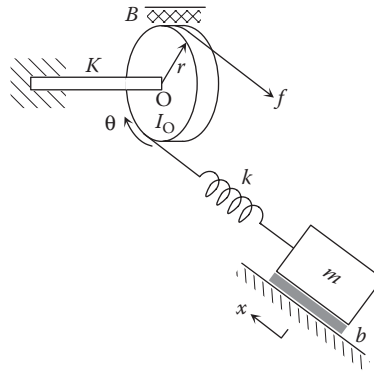


FIGURE 5.90
Problem 3.

3. Consider the mechanical system shown in [Figure 5.90](#). A disk–shaft system is connected to a block of mass m through a translational spring of stiffness k . The elasticity of the shaft and the fluid coupling are modeled as a torsional spring of stiffness K and a torsional viscous damper of damping coefficient B , respectively. The radius of the disk is r and its mass moment of inertia about point O is I_O . Assume that the friction between the block and the horizontal surface cannot be ignored and is modeled as a translational viscous damper of damping coefficient b . The input to the system is the force f . Draw the necessary free-body diagram and the kinematic diagram, and derive the equations of motion.
4. Consider the mechanical system shown in [Figure 5.91](#), where the motion of the rod is small angular rotation. When $\theta = 0$ and $f = 0$, the deformation of each spring is zero and the system is at static equilibrium. Assume that the friction between the block of mass m_1 and the horizontal surface cannot be ignored and is modeled as a translational viscous damper of damping coefficient b .

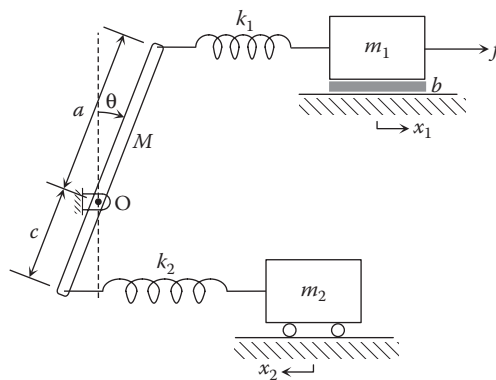
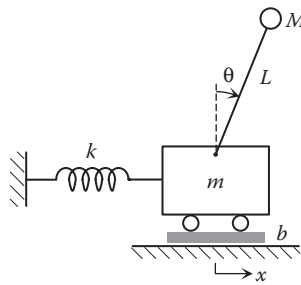
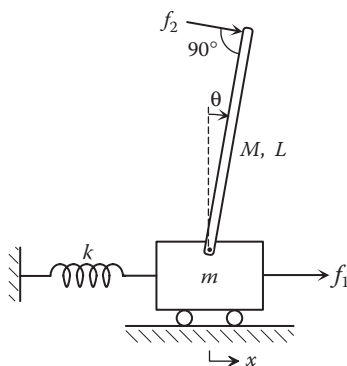


FIGURE 5.91
Problem 4.

**FIGURE 5.92**

Problem 5.

- a. Assuming that $a > c > 0$, determine the mass moment of inertia of the rod about the pivot point O.
 - b. Draw the necessary free-body diagram and the kinematic diagram, and derive the equations of motion for small angles.
5. Consider the mechanical system shown in Figure 5.92, in which a simple pendulum is pivoted on a cart of mass m and is constrained to rotate in a vertical plane. The pendulum consists of a point mass M concentrated at the tip of a massless rod of length L . Assume that the friction between the cart and the horizontal surface cannot be ignored. Denote the displacement of the cart as x and the angular displacement of the pendulum as θ . Draw the necessary free-body diagram and kinematic diagram, and derive the equations of motion for small angles.
6. Consider the mechanical system shown in Figure 5.93. The inputs are the force f_1 applied to the cart and the force f_2 applied at the tip of the rod. The outputs are the displacement x of the cart and the angular displacement θ of the pendulum.
- a. Draw the free-body diagram and kinematic diagram, and derive the equations of motion for small angles.
 - b. Using the differential equation obtained in Part (a), determine the state-space representation.
7. For the mechanical system in Problem 2, use the energy method to derive the equation of motion.

**FIGURE 5.93**

Problem 6.

8. For the mechanical system in Problem 11 of Problem Set 5.3, use the energy method to derive the equation of motion.
9. Repeat Problem 13 of Problem Set 5.3, using Lagrange's equations.
10. Repeat Problem 14 of Problem Set 5.3, using Lagrange's equations.
11. Repeat Problem 5, using Lagrange's equations.
12. Repeat Problem 6, using Lagrange's equations.
13. A robot arm consists of rigid links connected by joints allowing the relative motion of neighboring links. The dynamic model for a robot arm can be derived using Lagrange's equations

$$\frac{d}{dt} \left(\frac{\partial T}{\partial \dot{\theta}_i} \right) - \frac{\partial T}{\partial \theta_i} + \frac{\partial V}{\partial \theta_i} = \tau_i, \quad i = 1, 2, \dots, n$$

where:

θ_i is the angular displacement of the i th joint

τ_i is the torque applied to the i th joint

n is the total number of joints

Consider a single-link planar robot arm, as shown in Figure 5.94. Use Lagrange's equations to derive the dynamic model of the robot arm. Assume that the motion of the robot arm is constrained in a vertical plane, and the joint angle varies between 0° and 360° .

14. Repeat Problem 13 for a two-link planar robot arm, as shown in Figure 5.95. Assume that the motion of the robot arm is constrained in a horizontal plane, and the joint angles vary between 0° and 360° .

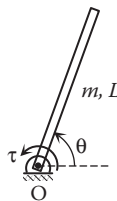


FIGURE 5.94
Problem 13.

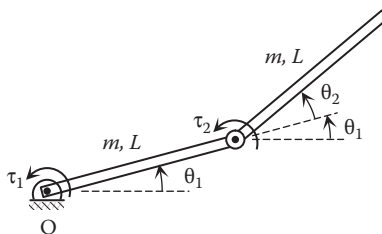


FIGURE 5.95
Problem 14.

5.5 Gear–Train Systems

Gear–train systems are important in many engineering applications. Figure 5.96a shows a pair of ideal gears, which are assumed to be rigid and meshed without backlash. A torque τ_1 produced by a motor is applied to gear 1, which rotates and causes gear 2 to rotate in the opposite direction. The radii of gear 1 and gear 2 are r_1 and r_2 , respectively. The relative sizes of the two gears result in a proportionality constant between the angular velocities of the respective shafts.

For the purpose of analysis, a free-body diagram for the rotational gear–train system must be drawn with care. It is convenient to visualize the gears as circles. Figure 5.96b shows the free-body diagram seen from the side of the input shaft. The two gears are tangent at the contact point and rotate without slipping. The torque applied to gear 1 causes an action force F onto gear 2 at the contact point. Because of Newton’s third law, gear 1 is subjected to a reaction force at the contact point. Use θ_1 and θ_2 to denote the respective angular displacements. The geometric constraint is

$$r_1\theta_1 = r_2\theta_2 \quad (5.52)$$

which can be rewritten as

$$\frac{\theta_2}{\theta_1} = \frac{r_1}{r_2} = N \quad (5.53)$$

where N is called the gear ratio, which also defines the relationship between the numbers of teeth on the two gears:

$$\frac{n_1}{n_2} = N \quad (5.54)$$

Note that the gear ratio N may be defined differently among authors. Differentiating Equation 5.53 gives

$$\frac{\omega_2}{\omega_1} = N \quad (5.55)$$

Equations 5.54 and 5.55 show that if the number of teeth on the output gear is larger than the number of teeth on the input gear, the input gear must rotate faster than the output

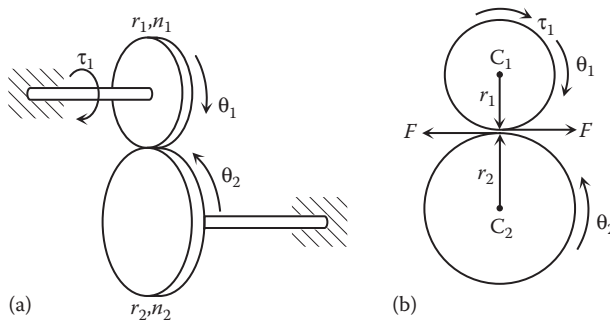


FIGURE 5.96

A gear–train system: (a) physical system and (b) free-body diagram.

gear. Thus, the gear pair is a speed reducer for $N < 1$. If the gears have negligible inertia or zero angular acceleration, and if the energy loss due to friction between the gear teeth can be neglected, the input work must be equal to the output work. Under these conditions, the output torque is greater than the input torque for a speed reducer.

To obtain the mathematical model of a gear–train system, the fundamental laws are still applied. We need to draw a free-body diagram and apply the moment equation for each gear. The derivation also requires a consideration of the geometric constraint.

Example 5.19: A Single-Degree-of-Freedom Gear–Train System

For the gear–train system shown in [Figure 5.96a](#), derive the differential equation of motion. The mass moments of inertia of the two gears about their respective fixed centers are I_{C1} and I_{C2} .

Solution

The free-body diagram is shown in [Figure 5.96b](#). The geometric constraint is

$$\frac{\theta_2}{\theta_1} = \frac{r_1}{r_2}$$

Applying the moment equation to each gear gives

$$\begin{aligned} + \curvearrowright: \sum M_{C1} &= I_{C1}\alpha \\ \tau_1 - r_1 F &= I_{C1}\ddot{\theta}_1 \end{aligned}$$

and

$$\begin{aligned} + \curvearrowright: \sum M_{C2} &= I_{C2}\alpha \\ r_2 F &= I_{C2}\ddot{\theta}_2 \end{aligned}$$

Combining the two equations and eliminating F yields

$$\tau_1 - \frac{r_1}{r_2} I_{C2} \ddot{\theta}_2 = I_{C1} \ddot{\theta}_1$$

Note that the angular displacements θ_1 and θ_2 are dependent through the geometric constraint. Thus, the gear–train in [Figure 5.96](#) is a single-degree-of-freedom system, which requires only one equation of motion in terms of only one coordinate. Assume that the equation is expressed in terms of θ_1 . Introducing the geometric constraint results in the equation of motion

$$\tau_1 - \frac{r_1}{r_2} I_{C2} \left(\frac{r_1}{r_2} \ddot{\theta}_1 \right) = I_{C1} \ddot{\theta}_1$$

simplified to

$$\left(I_{C1} + \frac{r_1^2}{r_2^2} I_{C2} \right) \ddot{\theta}_1 = \tau_1$$

which is the dynamics seen from the input side.

Example 5.20: A Single-Link Robot Arm

The mechanical model of a single-link robot arm driven by a motor can be represented as a gear–train system, as shown in [Figure 5.97](#), in which two rotational subsystems are coupled with a pair of gears with negligible inertia. The mass moments of inertia of the motor and the load are I_m and I , respectively. The coefficients of torsional viscous damping of the motor and the load are B_m and B , respectively. τ_m is the torque generated by the motor. Assume that the gear ratio is $N = r_1/r_2$. Derive the differential equation of motion in terms of the motor variable θ_m .

Solution

The free-body diagrams for the motor, the load, and the gear–train are shown in [Figure 5.98](#), where F represents the contact force between the two gears. The moments caused by the contact force on the motor and on the load are r_1F and r_2F , respectively. Applying the moment equation to the motor and the load gives

$$+\curvearrowright: \sum M_O = I_O \alpha$$

$$\tau_m - B_m \dot{\theta}_m - r_1 F = I_m \ddot{\theta}_m$$

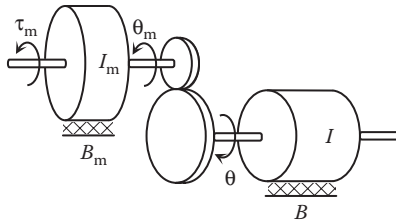
and

$$+\curvearrowright: \sum M_O = I_O \alpha$$

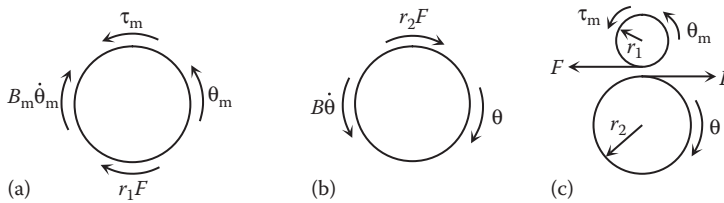
$$-B\dot{\theta} + r_2 F = I\ddot{\theta}$$

Solving for F from the equation for the load and substituting it into the equation for the motor results in

$$\tau_m - B_m \dot{\theta}_m - \frac{r_1}{r_2} (I\ddot{\theta} + B\dot{\theta}) = I_m \ddot{\theta}_m$$

**FIGURE 5.97**

The mechanical model of a single-link robot arm driven by a motor.

**FIGURE 5.98**

Free-body diagrams: (a) motor, (b) load, and (c) gear–train.

By the geometry of the gears,

$$\dot{\theta} = \frac{r_1}{r_2} \dot{\theta}_m = N \dot{\theta}_m$$

$$\ddot{\theta} = \frac{r_1}{r_2} \ddot{\theta}_m = N \ddot{\theta}_m$$

Substituting these into the previous differential equation gives

$$\tau_m - B_m \dot{\theta}_m - N^2(I \ddot{\theta}_m + B \dot{\theta}_m) = I_m \ddot{\theta}_m$$

which can be rearranged as

$$(I_m + N^2 I) \ddot{\theta}_m + (B_m + N^2 B) \dot{\theta}_m = \tau_m$$

The equation is in terms of the angular displacement of the motor.

PROBLEM SET 5.5

1. Repeat Example 5.19, and determine a mathematical model for the simple one-degree-of-freedom system shown in [Figure 5.96a](#) in the form of a differential equation of motion in θ_2 .
2. Repeat Example 5.20, and determine a mathematical model for the single-link robot arm shown in [Figure 5.97](#) in the form of a differential equation of motion in the load variable θ .
3. Consider the one-degree-of-freedom system shown in [Figure 5.99](#). The system consists of two gears of mass moments of inertia I_1 and I_2 and radii r_1 and r_2 , respectively. The applied torque on gear 1 is τ_1 . Assume that the gears are connected with flexible shafts, which can be approximated as two torsional springs of stiffness, K_1 and K_2 , respectively.
 - a. Draw the necessary free-body diagrams, and derive the differential equation of motion in θ_1 .
 - b. Using the differential equation obtained in Part (a), determine the transfer function $\Theta_2(s)/T_1(s)$.
 - c. Using the differential equation obtained in Part (a), determine the state-space representation with θ_2 as the output.

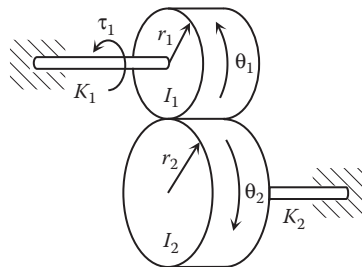


FIGURE 5.99
Problem 3.

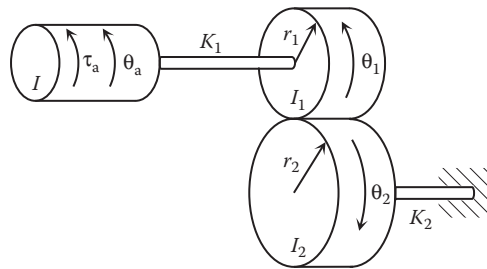


FIGURE 5.100
Problem 4.

4. Consider the gear-train system shown in [Figure 5.100](#). The system consists of a rotational cylinder and a pair of gears. The gear ratio is $N = r_1/r_2$. The applied torque on the cylinder is τ_a . Assume that the gears are connected with flexible shafts, which can be approximated as two torsional springs of stiffness, K_1 and K_2 , respectively.
 - a. Draw the necessary free-body diagrams, and derive the differential equations of motion.
 - b. Using the differential equations obtained in Part (a), determine the state-space representation. Use θ_a , θ_1 , ω_a , and ω_1 as the state variables, and use θ_2 and ω_2 as the output variables.
5. A three-degree-of-freedom gear-train system is shown in [Figure 5.101](#), which consists of four gears of moments of inertia I_1 , I_2 , I_3 , and I_4 . Gears 2 and 3 are meshed and their radii are r_2 and r_3 , respectively. Gears 1 and 2 are connected by a relatively long shaft, and gears 3 and 4 are connected in the same way. The shafts are assumed to be flexible, and can be approximated by torsional springs. The applied torque and load torque are τ_a and τ_l on gear 1 and gear 4, respectively. The gears are assumed to be rigid and have no backlash. Derive the differential equations of motion.
6. Repeat Problem 5. Assume that the shaft connecting gears 1 and 2 is relatively short and rigid.

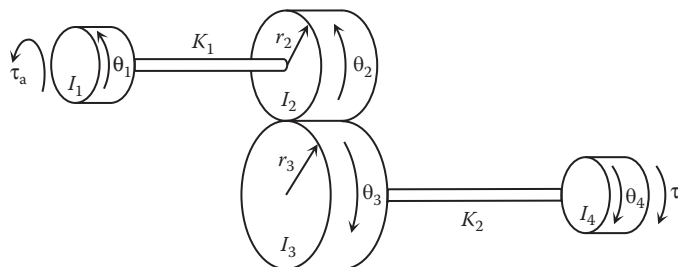


FIGURE 5.101
Problem 5.

5.6 System Modeling with Simulink and Simscape

Simulink is a block diagram environment for multidomain simulation and model-based design. It provides a graphical editor, block libraries, and solvers for modeling, simulating, and analyzing dynamic systems, and even connecting your model to hardware for real-time testing. Simscape is an extension of Simulink for modeling and simulating physical systems spanning mechanical, electrical, hydraulic, and other physical domains. The reader can refer to [Sections 1.6](#) and [1.7](#) for a brief introduction to Simulink and Simscape. In this section, more examples are included to illustrate mechanical system modeling with Simulink and Simscape. The mathematical models in all examples are given without a detailed derivation and the reader can derive them as an exercise.

5.6.1 Translational Systems

It is known that translational mechanical systems can be modeled as systems with interconnected mass, spring, and damper elements. The dynamics of such systems can be represented by ordinary differential equations, transfer functions, or in the state-space form. The common inputs are forces or displacements, and the common outputs are displacements, velocities, or accelerations.

To model a translational mechanical system with Simulink and Simscape, three parts of blocks are needed to simulate the input, the system itself, and the output. A Simulink diagram is constructed based on the system's mathematical representation, such as differential equation(s), transfer function(s), or the state-space form. Blocks in the Simulink libraries of Sources, Continuous, and Sinks are commonly used to construct the input, the system itself, and the output, respectively. While a Simscape block diagram is built just as you would assemble a physical system. Basic physical components like mass, spring, damper, etc. are available in the Simscape library of Translational Elements for representing one-dimensional translational motion. The blocks in the Simscape libraries of Mechanical Sources and Mechanical Sensors are used to generate inputs and output measurements.

Example 5.4 in [Section 5.2](#) shows how to build a Simulink or Simscape block diagram of a single-degree-of-freedom mass–spring–damper system with a force input. The mathematical model of the system is given by an ordinary differential equation. In this section, we will consider different numbers of degrees of freedom, different system inputs, and different system representations.

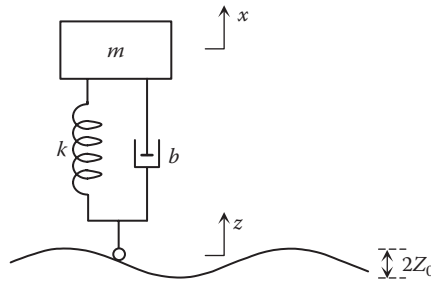
Example 5.21: A Single-Degree-of-Freedom Vehicle Model

The mass–spring–damper system shown in [Figure 5.102](#) represents a vehicle traveling on a rough road. Assume that the surface of the road can be approximated as a sine wave $z = Z_0 \sin(\omega t)$, where $Z_0 = 0.01$ m and $\omega = 3.5$ rad/s. The mathematical model of the system is given by an ordinary differential equation

$$m\ddot{x} + b\dot{x} + kx = b\dot{z} + kz$$

where $m = 3000$ kg, $b = 2000$ N·s/m, and $k = 50$ kN/m.

- Build a Simulink model of the system based on the mathematical representation and find the displacement output $x(t)$.
- Convert the ordinary differential equation to a transfer function and repeat Part (a). Assume zero initial conditions.
- Build a Simscape model of the physical system and find the displacement output $x(t)$.

**FIGURE 5.102**

The mechanical model of a vehicle traveling on a rough road.

Solution

- a. Solving for the highest derivative of the output x gives

$$\ddot{x} = \frac{1}{m}(kz + b\dot{z} - kx - b\dot{x})$$

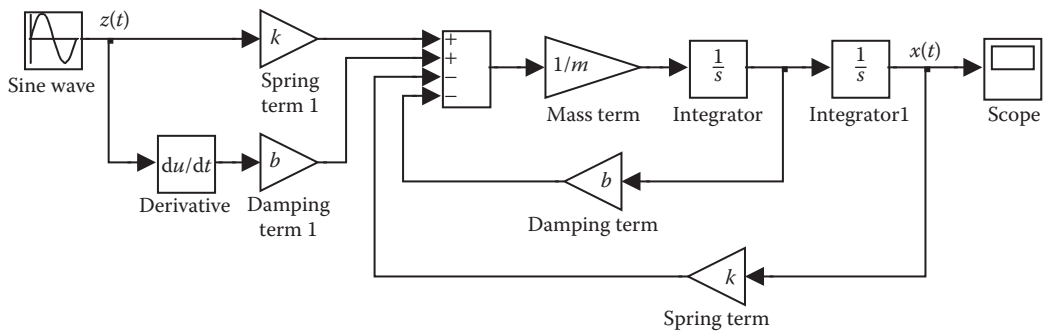
The corresponding Simulink block diagram is shown in [Figure 5.103](#). Note that the displacement input $z(t)$ is a sine function, which can be defined using a Sine Wave block available in the library of Sources. Double-click the block and type 0.01 for the Amplitude and 3.5 for the Frequency to define the input $z(t) = 0.01\sin(3.5t)$.

- b. The transfer function relating the input $z(t)$ to the output $x(t)$ is

$$\frac{X(s)}{Z(s)} = \frac{bs + k}{ms^2 + bs + k}$$

The Simulink block diagram built based on the transfer function is shown in [Figure 5.104](#), where a Transfer Fcn block is used to represent the vehicle system. Double-click the block and type [b k] for the Numerator coefficient and [m b k] for the Denominator coefficient to define the transfer function $X(s)/Z(s)$.

- c. The Simscape block diagram corresponding to the physical system is shown in [Figure 5.105](#), which can be created by following these steps:

**FIGURE 5.103**

Simulink block diagram built based on the ordinary differential equation.

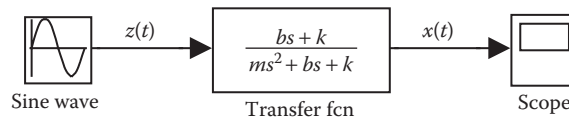


FIGURE 5.104
Simulink block diagram built based on the transfer function.

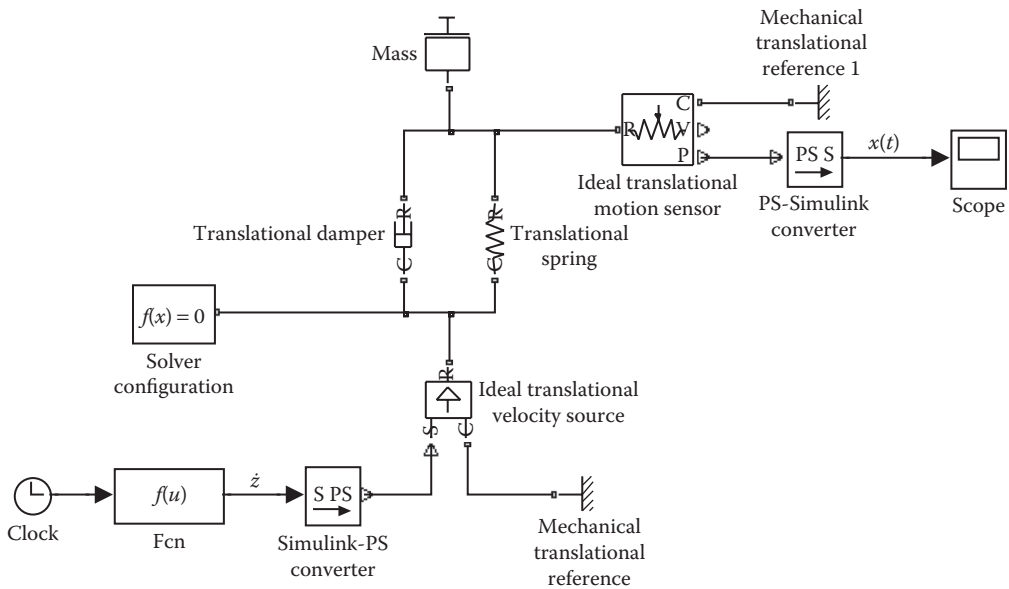
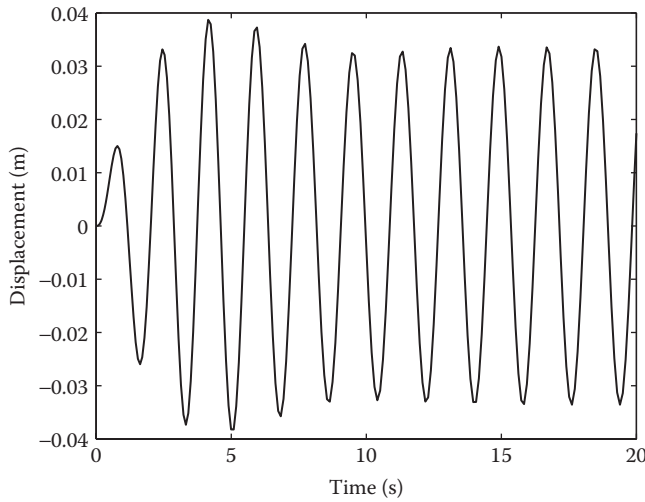


FIGURE 5.105
Simscape block diagram corresponding to Example 5.21.

1. Type `ssc_new` at the MATLAB Command window to open the main Simscape library and create a new model.
2. Open the library of Simscape/Foundation Library/Mechanical/Translational Elements and drag the Mass, Translational Damper, and Translational Spring into the model window. Double-click these blocks to define the parameters Mass, Damping coefficient, and Spring rate as m , b , and k .
3. To add the representation of the displacement input, open the library of Simscape/Foundation Library/Mechanical/Mechanical Sources and drag the Ideal Translational Velocity Source into the model window. Note that two types of inputs are available in Simscape for translational mechanical systems, and they are used to define either a force or a velocity input. Therefore, the displacement input function in this example must be converted to the velocity by taking the time derivative.
4. To add the sensor to measure the displacement of the mass, open the library of Simscape/Foundation Library/Mechanical/Mechanical Sensors and drag the Ideal Translational Motion Sensor into the model window.

**FIGURE 5.106**

Displacement output $x(t)$ of the vehicle model in Example 5.21.

5. Now open Simulink libraries to add the source and the scope. As mentioned in Step 3, taking the time derivative of $z(t)$ gives the velocity input function $\dot{z}(t) = 0.035\cos(3.5t)$. Open the library of Simulink/Sources to drag the Clock block and open the library of Simulink/User-Defined Functions to drag the Fun block. Double-click the Fun block and type $0.035*\cos(3.5*u)$ for the Expression, where u is the default name of the input to the Fun block, and here it represents the time t . Note that the Simulink-PS Converter and PS-Simulink Converter blocks are used to convert Simulink signals into physical signals or vice versa.
6. Orient the blocks and connect them, as shown in Figure 5.105.

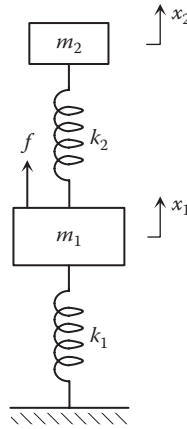
Define the values of the parameters m , b , and k in the MATLAB Command window. Run all simulations and the same curve, as shown in Figure 5.106, can be obtained, which is the resulting displacement output $x(t)$ of the vehicle due to the roughness of the road.

Example 5.22: A Two-Degree-of-Freedom Mass–Spring System

Consider the two-degree-of-freedom mass–spring system shown in Figure 5.107. The mass block m_1 and the spring k_1 represent a rotating machine, which is subjected to a harmonic disturbance force $f = 40\sin(7\pi t)$ N due to a rotating unbalanced mass. The mass block m_2 and the spring k_2 represent a vibration absorber (see Section 9.3 for more details), which is designed to reduce the displacement of the machine. The mathematical model of the system is given by a set of ordinary differential equations

$$m_1\ddot{x}_1 + (k_1 + k_2)x_1 - k_2x_2 = f$$

$$m_2\ddot{x}_2 - k_2x_1 + k_2x_2 = 0$$

**FIGURE 5.107**

A two-degree-of-freedom mass–spring system.

where $m_1 = 6$ kg, $k_1 = 6000$ N/m, $m_2 = 1.65$ kg, and $k_2 = 800$ N/m. Assume zero initial conditions.

- Build a Simulink model of the system based on the differential equations of motion and find the displacement outputs $x_1(t)$ and $x_2(t)$.
- Convert the ordinary differential equations to the state-space form and repeat Part (a).
- Build a Simscape model of the physical system and find the displacement outputs $x_1(t)$ and $x_2(t)$.

Solution

- Solving for the highest derivatives of the output x_1 and x_2 , respectively, gives

$$\ddot{x}_1 = \frac{1}{m_1} [f - (k_1 + k_2)x_1 + k_2x_2]$$

$$\ddot{x}_2 = \frac{1}{m_2} (k_2x_1 - k_2x_2)$$

The corresponding Simulink block diagram is shown in [Figure 5.108](#), where four Integrator blocks are included to obtain the signals $\dot{x}_1$, x_1 , $\dot{x}_2$, and x_2 . A Sine Wave block is used to model the harmonic disturbance force $f = 40\sin(7\pi t)$ N. Double-click the block and type 40 for the Amplitude and $7*\pi$ for the Frequency to define the disturbance input.

- Define the state, the input, and the output vectors as

$$\mathbf{x} = \begin{Bmatrix} x_1 \\ x_2 \\ \dot{x}_1 \\ \dot{x}_2 \end{Bmatrix} = \begin{Bmatrix} x_1 \\ x_2 \\ \dot{x}_1 \\ \dot{x}_2 \end{Bmatrix}, \quad u = f, \quad \mathbf{y} = \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix}$$

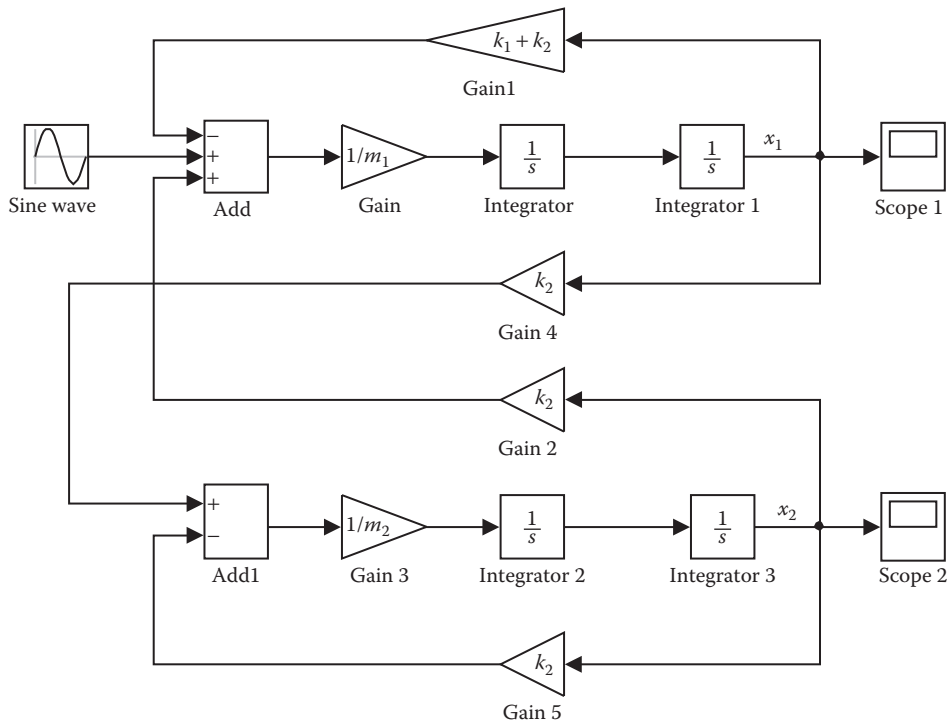


FIGURE 5.108
Simulink block diagram built based on the ordinary differential equations.

The state-space representation is

$$\begin{Bmatrix} \dot{x}_1 \\ \dot{x}_2 \\ \dot{x}_3 \\ \dot{x}_4 \end{Bmatrix} = \begin{bmatrix} 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 1 \\ -\frac{k_1+k_2}{m_1} & \frac{k_2}{m_1} & 0 & 0 \\ \frac{k_2}{m_2} & -\frac{k_2}{m_2} & 0 & 0 \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \\ x_3 \\ x_4 \end{Bmatrix} + \begin{bmatrix} 0 \\ 0 \\ \frac{1}{m_1} \\ 0 \end{bmatrix} u, \quad \mathbf{y} = \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \\ x_3 \\ x_4 \end{Bmatrix}$$

The Simulink block diagram built based on the state-space form is shown in [Figure 5.109](#), in which a State-Space block is used to represent the mass-spring system. Same as Part (a), a Sine Wave block is used to model the input. Double-click the State-Space block and define the matrices **A**, **B**, **C**, and **D**, which is a 2×1 zero vector. The parameter of Initial conditions is a 4×1 zero vector. The bar-shaped block in [Figure 5.109](#) is called Demux, which can be found in the library of Signal Routing and is used to split the vector signal **y** into two signals x_1 and x_2 .

- c. The Simscape block diagram corresponding to the physical system is shown in [Figure 5.110](#), which can be created by following the steps similar to those in Example 5.21.

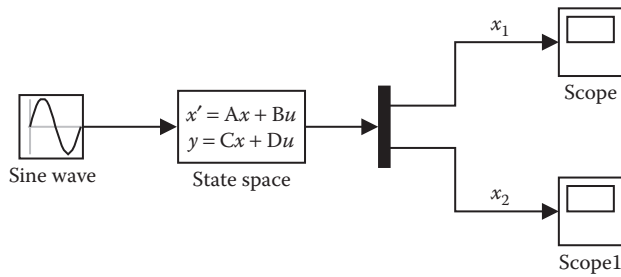


FIGURE 5.109
Simulink block diagram built based on the state-space form.

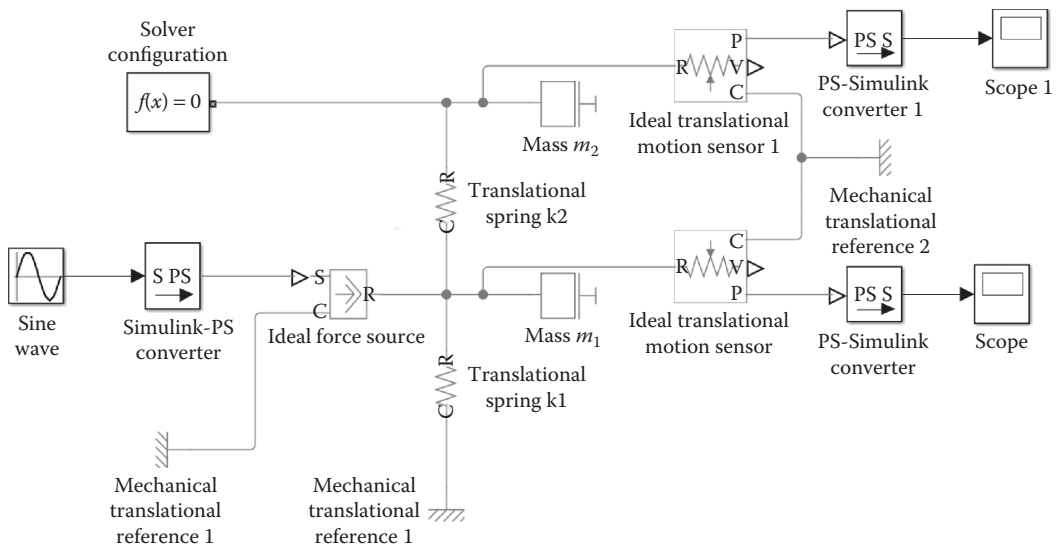


FIGURE 5.110
Simscape block diagram corresponding to Example 5.22.

Define the values of the parameters m_1 , m_2 , k_1 , and k_2 in the MATLAB Command window. Run all simulations and the same curves, as shown in [Figure 5.111](#), will be obtained, which are the resulting displacement outputs $x_1(t)$ and $x_2(t)$ due to the harmonic disturbance force input.

The aforementioned two examples and Example 5.4 demonstrate basic modeling techniques with Simulink and Simscape. Since a Simulink block diagram of a dynamic system is created based on the system's mathematical model, the modeling techniques discussed in this section can be applied to other types of dynamic systems represented as ordinary differential equations, transfer functions, or in the state-space form. However, a Simscape model describes the physical structure of a dynamic system rather than the underlying mathematics. It is obvious that Simscape diagrams look different for different types of dynamic systems. Therefore, [Sections 6.6](#) and [7.4](#) will focus on Simscape modeling in various physical domains, such as rotational mechanical, electrical, hydraulic, and thermal.

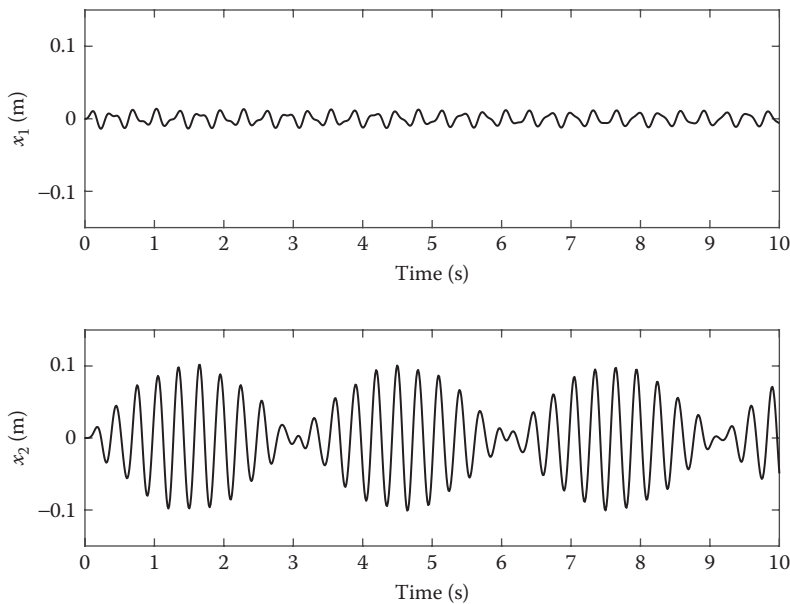


FIGURE 5.111
Displacement output $x(t)$ of the vehicle model in Example 5.22.

5.6.2 Rotational Systems

The blocks in the library of Simscape\Foundation Library\Mechanical can be categorized into two classes: translational and rotational. Basic elements such as inertia, rotational damper, and rotational spring available in the library of Rotational Elements are used to model a rotational mechanical system. Two source blocks, Ideal Torque Source and Ideal Angular Velocity Source, in the library of Mechanical Sources are used to generate inputs. Also, two sensors, Ideal Torque Sensor and Ideal Rotational Motion Sensor, in the library of Mechanical Sensors are used to output measurements.

Example 5.23: A Single-Degree-of-Freedom Rotational Mass–Spring–Damper System

Consider the simple disk–shaft system in Example 5.8. A single-degree-of-freedom rotational mass–spring–damper system can be used to approximate the dynamic behavior of the disk–shaft system. The parameter values are $I = 0.01 \text{ kg} \cdot \text{m}^2$, $B = 1.15 \text{ N} \cdot \text{m} \cdot \text{s}/\text{rad}$, and $K = 4150 \text{ N} \cdot \text{m}/\text{rad}$.

- Assume that a torque $\tau = 10\sin(600t)$ is acting on the disk, which is initially at rest. Build a Simscape model of the physical system and find the angular displacement output $\theta(t)$.
- Assuming that the external torque is $\tau = 0$ and the initial angular displacement is $\theta(0) = 0.1 \text{ rad}$, find the angular displacement output $\theta(t)$.

Solution

- The Simscape block diagram and the angular displacement output of the system are shown in Figures 5.112 and 5.113, respectively. Comparing

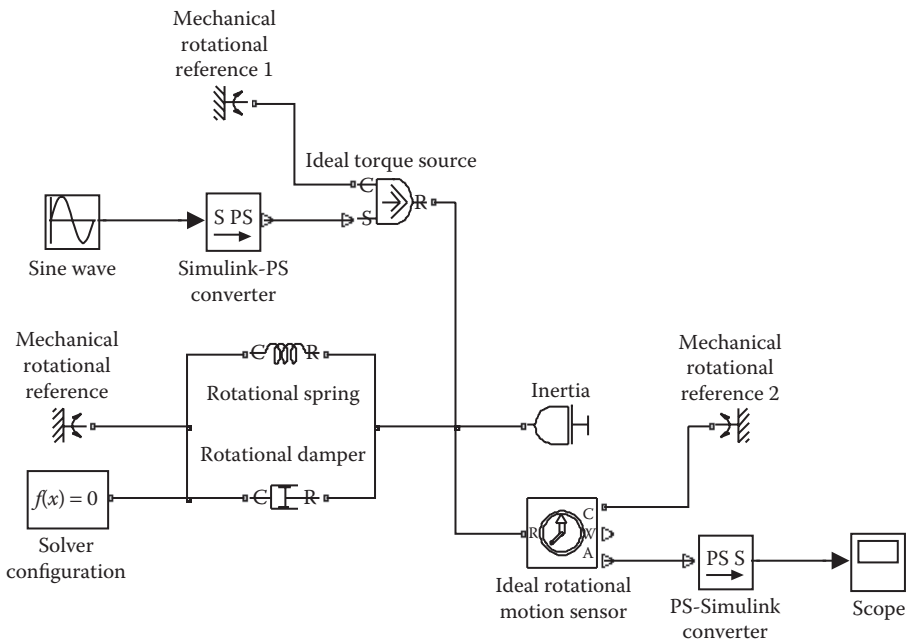


FIGURE 5.112
Simscape block diagram of the physical system in Example 5.23.

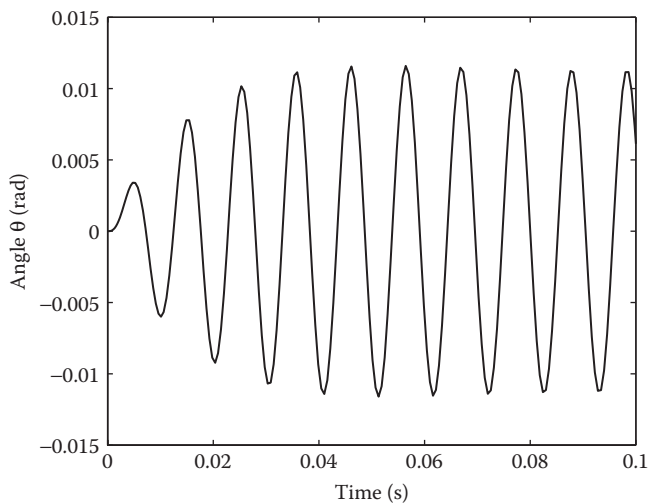
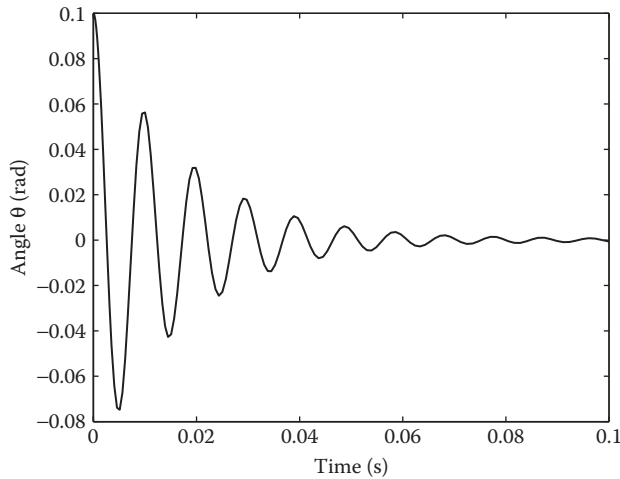


FIGURE 5.113
Angular displacement output $\theta(t)$ of the system in Example 5.23 with a torque input.

Figure 5.112 with Figure 5.31, which is the Simscape model of a single-degree-of-freedom translational mass-spring-damper system, reveals the similarity of these two Simscape diagrams. The main difference is that the blocks in this example are all related to rotational motion instead of translational motion.

**FIGURE 5.114**

Angular displacement output $\theta(t)$ of the system in Example 5.23 with a non-zero initial condition.

- b. The Simscape model in Figure 5.111 can also be used to simulate the system in Part (b). To specify a zero external torque, we can either define the Amplitude of the Sine Wave as 0 or delete the blocks related to input generation, including Sine Wave, Simulink-PS Converter, Ideal Torque Source, and Mechanical Rotational Reference blocks. To specify a non-zero initial angle, double-click the Rotational Spring block, type 0.1 for the Initial deformation, and choose the unit as rad. This implies that the spring is initially twisted by 0.1 rad. Also, double-click the Ideal Rotational Motion Sensor block, type 0.1 for the Initial angle, and choose the unit as rad. The corresponding angular displacement of the system is shown in Figure 5.114.

Example 5.24: A Single-Link Robot Arm with a Gearbox

Consider the single-link robot arm in Example 5.20. It is driven by a DC motor through a gearbox with a gear ratio $N = 1/70$. The mass moments of inertia of the motor and the load are I_m and I , respectively. The coefficients of torsional viscous damping of the motor and the load are B_m and B , respectively. τ_m is the torque generated by the motor. The parameter values are $I_m = 3.87 \times 10^{-7} \text{ kg} \cdot \text{m}^2$, $I = 0.0016 \text{ kg} \cdot \text{m}^2$, $B_m = 0 \text{ N} \cdot \text{m} \cdot \text{s}/\text{rad}$, and $B = 0.004 \text{ N} \cdot \text{m} \cdot \text{s}/\text{rad}$. Assume that the robot arm is initially at rest and the motor torque is $\tau_m(t) = 0.003u(t)$, where $u(t)$ is the unit-step function with a step time of 0 seconds. Build a Simscape model of the physical system and find the angular velocity output $\dot{\theta}(t)$.

Solution

The Simscape model of the system is shown in Figure 5.115. To model the robot arm, two Inertia blocks are included, one for the motor and the other for the load. Since the damping of the motor can be ignored, one Rotational Damper block is included to represent the damping of the load. The shaft of the motor and that of the load are coupled through a gearbox. Double-click the Gear Box block and type 70 (not 1/70) for the Gear ratio, which in Simscape is defined as the ratio of the input shaft angular velocity to that of the output shaft. This is the reciprocal of the gear ratio defined in Equation 5.55. The output of port W of the Ideal Rotation Motion Sensor is the corresponding angular velocity $\dot{\theta}(t)$, which is shown in Figure 5.116.

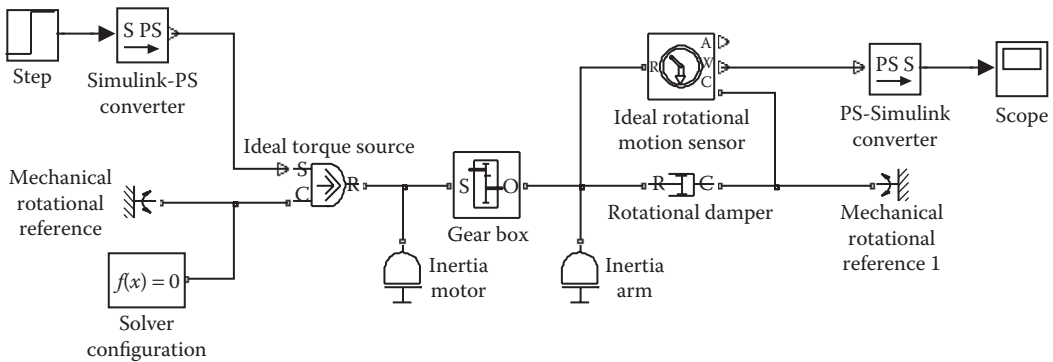


FIGURE 5.115
Simscape block diagram of the single-link robot arm with a gearbox.

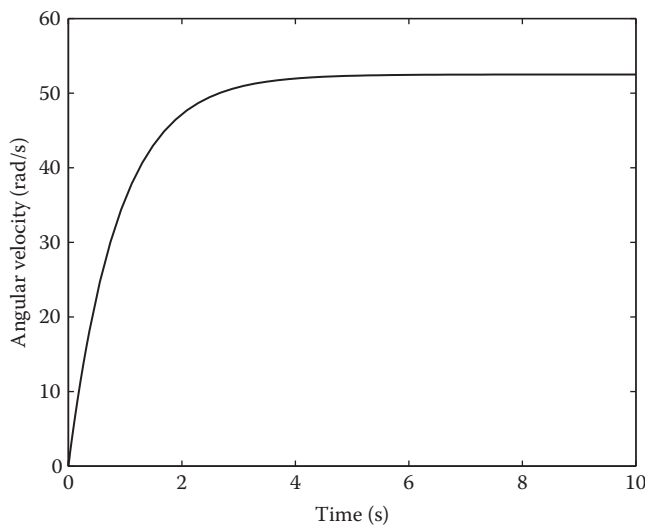


FIGURE 5.116
Angular velocity output $\dot{\theta}(t)$ of the robot arm in Example 5.24.

Note that only the Simscape block diagrams are given in the previous two examples. The corresponding Simulink modeling is left to the reader as an exercise.

PROBLEM SET 5.6

1. Consider the mass–spring–damper system in Problem 6 of Problem Set 5.2. Assume that the force acting on the mass block is a unit-impulse function with a magnitude of 100 N and a duration of 0.1 sec. The parameter values are $m = 500$ kg, $b = 250$ N·s/m, and $k = 200,000$ N/m.
 - a. Build a Simulink model based on the differential equation of motion of the system and find the displacement output $x(t)$.
 - b. Build a Simscape model of the physical system and find the displacement output $x(t)$.

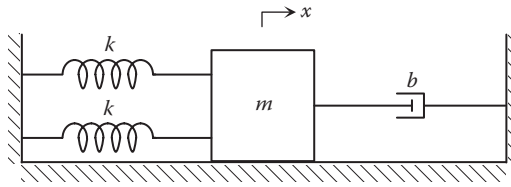


FIGURE 5.117
Problem 2.

2. Repeat Problem 1 for the mass–spring–damper system shown in Figure 5.117, in which the origin of the coordinate x is set at equilibrium. Assume that $x(0) = 0.1$ m and $\dot{x}(0) = 0$ m/s. The parameter values are $m = 20$ kg, $b = 125$ N s/m, and $k = 400$ N/m.
3. Consider the two-degree-of-freedom mass–spring system shown in Figure 5.118. The parameter values are $m_1 = m_2 = 5$ kg, $k_1 = 2000$ N/m, and $k_2 = 4000$ N/m. Assume that, initially, $x(0) = [0 \ 0]^T$ and $\dot{x}(0) = [1 \ 0]^T$.
 - a. Build a Simulink model based on the differential equations of motion of the system and find the displacement outputs $x_1(t)$ and $x_2(t)$.
 - b. Build a Simscape model of the physical system and find the displacement outputs $x_1(t)$ and $x_2(t)$.
4. Repeat Problem 3 for the two-degree-of-freedom quarter-car model in Example 5.5. Assume that the surface of the road can be approximated as a sine wave $z = Z_0 \sin(2\pi vt/L)$, where $Z_0 = 0.01$ m, $L = 10$ m, and the speed $v = 20$ km/h. If the car moves at a speed of 100 km/h, rerun the simulations and compare the results with those obtained in the case of 20 km/h. Ignore the control force f for both cases.
5. Consider the disk–shaft system in Problem 2 of Problem Set 5.3. The system is approximated as a single-degree-of-freedom rotational mass–spring system, where $m = 10$ kg, $r = 0.05$ m, $B = 1$ N · m · s/rad, and $K = 1000$ N · m/rad.
 - a. Assume that a torque $\tau = 50u(t)$ N · m is acting on the disk, which is initially at rest. Build a Simscape model of the physical system and find the angular displacement output $\theta(t)$.
 - b. Assuming that the external torque is $\tau = 0$ and the initial angular displacement is $\theta(0) = 0.1$ rad, find the angular displacement output $\theta(t)$.

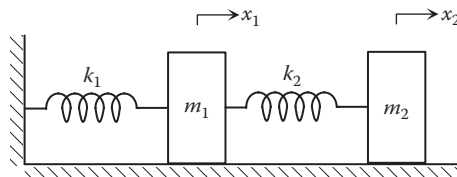


FIGURE 5.118
Problem 3.

6.  Consider the pendulum-bob system in Problem 5 of Problem Set 5.3. The parameter values are $m = 0.1$ kg, $M = 1.2$ kg, $L = 0.6$ m, and $B = 0.25$ N · s/m. The initial angular displacement is $\theta(0) = 0.1$ rad and the initial angular velocity is $\dot{\theta} = 0.1$ rad/s.
- Build a Simulink block diagram based on the nonlinear mathematical model of the system and find the angular displacement output $\theta(t)$.
 - Build a Simscape model of the nonlinear physical system and find the angular displacement output $\theta(t)$.

5.7 Summary

This chapter was devoted to modeling mechanical systems. Because real systems are usually quite complicated, simplifying assumptions must be made to reduce the system to an idealized model consisting of interconnected mass, damper, and spring elements. The relations between the external forces or moments applied to the elements and the associated element variables are given by

- | | | |
|-----------------------|-----------------------------|---|
| • Mass translation: | $f = m\ddot{x}$ | rotation about fixed O: $\tau = I_O\ddot{\theta}$ |
| • Damper translation: | $f = b\dot{x}_{\text{rel}}$ | rotation: $\tau = B\dot{\theta}_{\text{rel}}$ |
| • Spring translation: | $f = kx_{\text{rel}}$ | rotation: $\tau = K\theta_{\text{rel}}$ |

Here, the spring force is dependent on the relative displacement between the two ends of the spring, and the damping force depends on the relative velocity between the two ends of the damper.

For a system of interconnected mechanical elements, the dynamic equations of motion can be obtained by applying Newton's second law or the moment equation (or both). The number of equations of motion is determined by the number of degrees of freedom of the system. The number of degrees of freedom of a dynamic system is defined as the number of independent generalized coordinates that specify the configuration of the system. The static equilibrium position of a mechanical system is usually chosen as the coordinate origin. This choice can simplify the equation of motion by eliminating static forces.

To apply Newton's second law or the moment equation to a mechanical system, it is useful to draw a free-body diagram for each mass in the system, showing all external forces or moments. The noninput forces can be described in terms of displacements or velocities, using the expressions associated with the basic spring or damper elements. Drawing correct free-body diagrams is the most important step in analyzing mechanical systems, using the force/moment approach.

In this chapter, we were mainly concerned with the modeling of mechanical systems in plane motion, which involves translations along the x and y directions and rotation about one axis perpendicular to the x - y plane. Newton's second law is used for modeling translational mechanical systems, whereas the moment equation is used for rotational mechanical systems. Newton's second law and the moment equation are used together for mixed translational and rotational systems.

The general form of Newton's second law for a rigid body (or a particle) is given by

$$\sum \mathbf{F} = m\mathbf{a}_C$$

or

$$\sum F_x = ma_{Cx}, \quad \sum F_y = ma_{Cy}$$

For a rigid body rotating about a fixed axis through point O, the moment equation is given by

$$\sum M_O = I_O \alpha$$

If the axis of rotation is not fixed, the moment equation can be set about the mass center C as

$$\sum M_C = I_C \alpha$$

or an arbitrary point P as

$$\sum \mathbf{M}_P = I_P \alpha + m \mathbf{r}_{C/P} \times \mathbf{a}_P$$

which is equivalent to

$$\sum M_P = I_C \alpha + M_{\text{eff_mac}}$$

The symbol $M_{\text{eff_mac}}$ represents the effective moment caused by the fictitious force $m\mathbf{a}_C$.

The mass moments of inertia for some rigid bodies with common shapes were given in [Table 5.1](#), in which all masses are assumed to be uniformly distributed and all the axes of rotation pass through the mass centers. If the axis of rotation does not coincide with the axis through the mass center, but is parallel to it, the parallel-axis theorem can be applied to obtain the corresponding moment of inertia,

$$I = I_C + md^2$$

where d is the distance between the two parallel axes.

For a system of multiple masses, the force equations become

$$\sum F_x = \sum_{i=1}^n m_i (a_{Ci})_x, \quad \sum F_y = \sum_{i=1}^n m_i (a_{Ci})_y$$

and the moment equation becomes

$$\sum M_P = \sum_{i=1}^n I_{Ci} \alpha_i + \sum_{i=1}^n M_{\text{eff_m}i\mathbf{a}Ci}$$

The force/moment approach is based on Newtonian mechanics. An alternative way of obtaining the system's equations of motion is to use the energy method based on analytical mechanics. For a single-degree-of-freedom mass-spring system with negligible friction and damping, the principle of conservation of energy states that

$$\frac{d}{dt}(T + V) = 0$$

The kinetic energy of a rigid body in plane motion can be separated into two parts: (1) the kinetic energy associated with the translational motion of the mass center C of the body

and (2) the kinetic energy associated with the rotation of the body about C. Expressions for the kinetic energy of a rigid body in plane motion are given as

- Translation only: $T = \frac{1}{2}mv^2$
- Rotation about a fixed point O: $T = \frac{1}{2}I_O\omega^2$
- Mixed translation and rotation: $T = \frac{1}{2}mv_C^2 + \frac{1}{2}I_C\omega^2$

The potential energy includes

- Gravitational potential energy: $V_g = mgh$
- Elastic potential energy: $V_e = \frac{1}{2}kx^2$ or $V_e = \frac{1}{2}K\theta^2$

For an n -degree-of-freedom system, n independent equations of motion can be derived using Lagrange's formulation. One of the forms of Lagrange's equations for a conservative system is

$$\frac{d}{dt}\left(\frac{\partial T}{\partial \dot{q}_i}\right) - \frac{\partial T}{\partial q_i} + \frac{\partial V}{\partial q_i} = 0, \quad i = 1, 2, \dots, n$$

where:

q_i is the i th generalized coordinate

n is the total number of independent generalized coordinates

For a nonconservative system, the previous equation is changed to

$$\frac{d}{dt}\left(\frac{\partial T}{\partial \dot{q}_i}\right) - \frac{\partial T}{\partial q_i} + \frac{\partial V}{\partial q_i} = Q_i, \quad i = 1, 2, \dots, n$$

where Q_i is the i th generalized force. In contrast to the force/moment approach with Newton's second law and the moment equation, Lagrange's equations do not require a free-body diagram. However, velocity analysis is essential to Lagrange's equations.

Review Problems

1. Determine the equivalent spring constant for the system shown in [Figure 5.119](#).
2. Determine the equivalent spring constant for the system shown in [Figure 5.120](#).
3. Consider the system shown in [Figure 5.121](#), in which a mass-spring system is hung from the middle of a massless beam. Assume that the beam can be modeled as a spring and the equivalent stiffness at the mid-span is $192EI_A/L^3$, where E is the modulus of elasticity of beam material and I_A is the area moment of inertia about the beam's longitudinal axis.
 - a. Derive the differential equation of motion for the system.
 - b. Using the differential equation obtained in Part (a), determine the transfer function $X(s)/F(s)$. Assume that the initial conditions are $x(0) = 0$ and $\dot{x}(0) = 0$.
 - c. Using the differential equation obtained in Part (a), determine the state-space representation. Assume that the output is the displacement x of the mass.

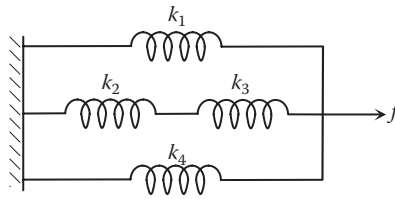


FIGURE 5.119
Problem 1.

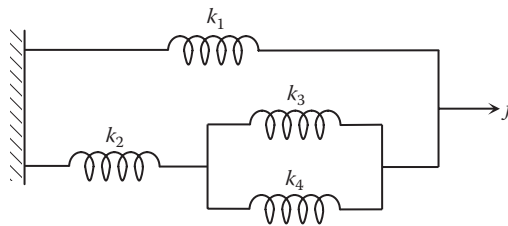


FIGURE 5.120
Problem 2.

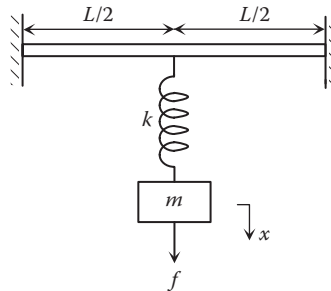


FIGURE 5.121
Problem 3.

4. An accelerometer attached to an object can be modeled as a mass–damper–spring system, as shown in [Figure 5.122](#). Denote the displacement of the mass relative to the object, the absolute displacement of the mass, and the absolute displacement of the object as $x(t)$, $y(t)$, and $z(t)$, respectively, where $x(t) = y(t) - z(t)$ and $x(t)$ is measured electronically.
 - a. Draw the necessary free-body diagram and derive the differential equation in terms of $x(t)$.
 - b. Using the differential equation obtained in Part (a), determine the transfer function $X(s)/Z(s)$. Assume that the initial conditions are $x(0) = 0$ and $\dot{x}(0) = 0$.
 - c. Using the differential equation obtained in Part (a), determine the state-space representation. The input is the absolute displacement of the object $z(t)$ and the output is the displacement of the mass relative to the object $x(t)$.

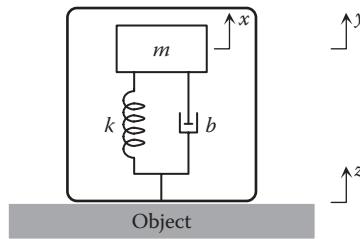


FIGURE 5.122
Problem 4.

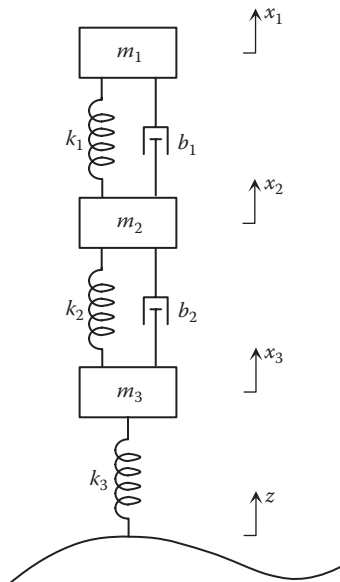
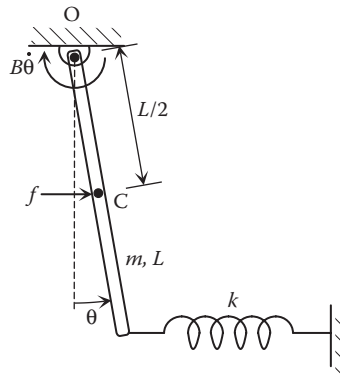


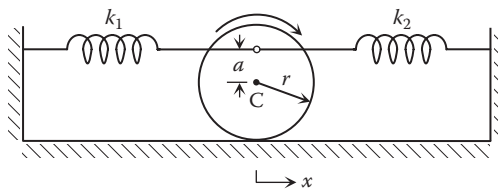
FIGURE 5.123
Problem 5.

5. Consider a quarter-car model shown in [Figure 5.123](#), where m_1 is the mass of the seats including passengers, m_2 is the mass of one-fourth of the car body, and m_3 is the mass of the wheel–tire–axle assembly. The spring k_1 represents the elasticity of the seat supports, k_2 represents the elasticity of the suspension, and k_3 represents the elasticity of the tire. $z(t)$ is the displacement input due to the surface of the road. Draw the necessary free-body diagrams and derive the differential equations of motion. Write the differential equations of motion in the second-order matrix form.
6. The system shown in [Figure 5.124](#) consists of a uniform rod of mass m and length L and a translational spring of stiffness k at the rod's tip. The friction at the joint O is modeled as a damper with coefficient of torsional viscous damping B . The input is the force f and the output is the angle θ . The position $\theta = 0$ corresponds to the static equilibrium position when $f = 0$.
 - a. Draw the necessary free-body diagram and derive the differential equation of motion for small angles θ .

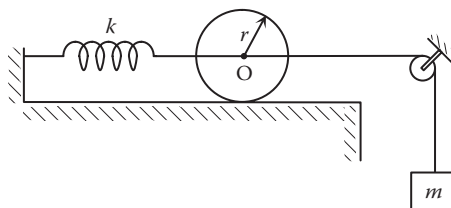
**FIGURE 5.124**

Problem 6.

- b. Using the linearized differential equation obtained in Part (b), determine the transfer function $\Theta(s)/F(s)$. Assume that the initial conditions are $\theta(0) = 0$ and $\dot{\theta}(0) = 0$.
 - c. Using the differential equation obtained in Part (b), determine the state-space representation.
7. Consider the system shown in Figure 5.125. Assume that a cylinder of mass m rolls without slipping. Draw the necessary free-body diagram and derive the differential equation of motion for small angles θ .
 8. The pulley of mass M shown in Figure 5.126 has a radius of r . The mass moment of inertia of the pulley about the point O is I_O . A translational spring of stiffness k and a block of mass m are connected to the pulley, as shown. Assume that the pulley rolls without slipping. Derive the equation of motion by using (a) the force/moment approach and (b) the energy approach.

**FIGURE 5.125**

Problem 7.

**FIGURE 5.126**

Problem 8.

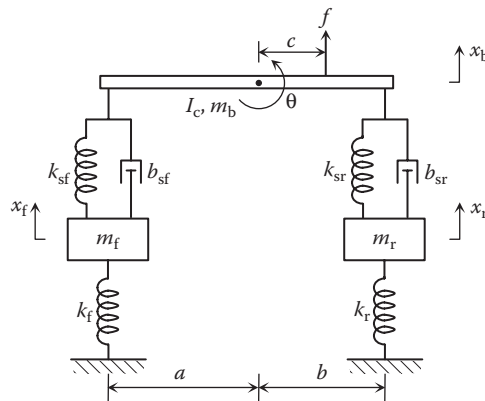


FIGURE 5.127
Problem 9.

9. Consider a half-car model shown in [Figure 5.127](#), in which I_c is the mass moment of inertia of the car body about the pitch axis, m_b is the mass of the car body, m_f is the mass of the front wheel-tire-axle assembly, and m_r is the mass of the rear wheel-tire-axle assembly. Each of the front and rear wheel-tire-axle assemblies is represented by a mass-spring-damper system. The input is the force f , and the car undergoes vertical and pitch motion. Derive the equations of motion by using the force/moment approach.
10. Consider the mechanical system shown in [Figure 5.128](#), in which a uniform rod of mass m and length L is attached to a massless rigid link of equal length. Assume that the system is constrained to move in a vertical plane. Denote the angular displacement of the link as θ_1 and the angular displacement of the rod as θ_2 . Derive the equations of motion for small angles by using (a) the force/moment approach and (b) the energy approach.
11. Consider the mechanical system shown in [Figure 5.129](#), in which a block of mass m is connected to a spring of stiffness k , and it can slide without friction inside a vertical tube. Assume that the tube pivots at the joint O , and it can be approximated as a uniform rod of mass M and length L . The unstretched length of the spring is $L/2$. Denote the angular displacement of the tube as θ and the relative displacement of the block with respect to the tube as x . Derive the nonlinear equations of motion by using (a) the energy approach and (b) the force/moment approach.

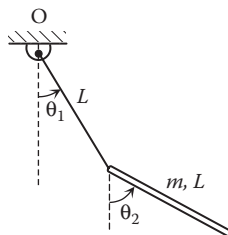


FIGURE 5.128
Problem 10.

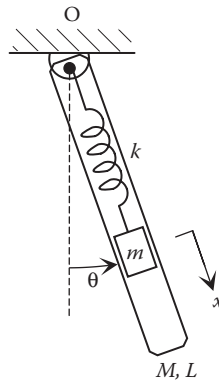


FIGURE 5.129
Problem 11.

12. A rack and pinion is a pair of gears that converts rotational motion into translation. As shown in [Figure 5.130](#), a torque τ is applied to the shaft. The pinion rotates and causes the rack to translate. The mass moment of inertia of the pinion is I and the mass of the rack is m . Draw the free-body diagram and derive the differential equation of motion.
13.  Consider the mass-spring-damper system shown in Problem 9 of Problem Set 5.2, in which the cam and follower impart a displacement $z(t)$ in the form of a periodic sawtooth function ([Figure 5.131](#)) to the lower end of the system. The values of the system parameters are $m = 12$ kg, $b = 200$ N · s/m, $k_1 = 4000$ N/m, and $k_2 = 2000$ N/m.
 - a. Build a Simulink model based on the differential equation of motion of the system and find the displacement output $x(t)$.
 - b. Build a Simscape model of the physical system and find the displacement output $x(t)$.
14. Case study

Consider the ball and beam system [10] shown in [Figure 5.132](#), in which I and I_b are the mass moments of inertia of the beam and the ball, respectively, m is the mass of the ball, and r is the radius of the ball. The beam is made to rotate in a vertical plane by applying a torque τ at the center of rotation and the ball is free to roll

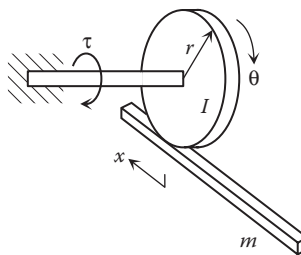


FIGURE 5.130
Problem 12.

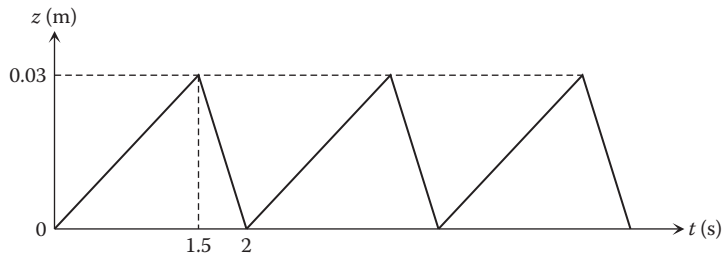


FIGURE 5.131
Problem 13.

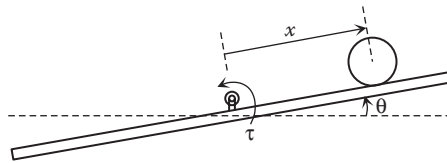


FIGURE 5.132
Problem 14.

(with one degree of freedom) along the beam. It is assumed that the ball remains in contact with the beam and that the rolling occurs without slipping. The system parameters are $I = 0.02 \text{ kg} \cdot \text{m}^2$, $I_b = 2 \times 10^{-6} \text{ kg} \cdot \text{m}^2$, $m = 0.05 \text{ kg}$, and $r = 0.01 \text{ m}$.

- Choosing the beam angle θ and the ball position x as generalized coordinates, derive the nonlinear equations of motion of the system by using the Lagrange's equations.
- Choosing θ , $\dot{\theta}$, x , and $\dot{x}$ as state variables and τ as the input, find the state variable equations of the nonlinear system.
- Assuming small angular motion, find the state-space form for the linearized system (with θ and x as outputs).
- Build two Simulink block diagrams, one using the differential equations obtained in Part (a) and the other using the state-space form found in Part (c). Compare the beam angle $\theta(t)$ and the ball displacement $x(t)$ when the torque applied to the system is a pulse function, $\tau(t) = 1 \text{ Nm}$ for $1 \leq t \leq 2 \text{ s}$.



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6

Electrical, Electronic, and Electromechanical Systems

Many engineering systems have electrical, electronic, or electromechanical subsystems as important components, such as power supplies, motors, sensors, and controllers. In this chapter, we discuss the modeling techniques for these systems. We first introduce the fundamentals of electrical elements, which include resistors, inductors, and capacitors. The two main physical laws, Kirchhoff's voltage law and Kirchhoff's current law, are then reviewed and applied to develop mathematical models of electrical circuits. For electronic systems, we study simple operational amplifiers (op-amps), and present the op-amp equation, which is useful for obtaining models of amplifiers. This is followed by the modeling of electromechanical systems. The coupling between electrical and mechanical subsystems is established and applied to motor modeling. To simplify the modeling of electrical systems, the concept of impedance is introduced, which provides an alternative way of obtaining mathematical models of systems. The chapter concludes with simulation of electrical, electronic, and electromechanical systems by using MATLAB®, Simulink®, and Simscape™ computer tools.

6.1 Electrical Elements

Electrical systems, or electrical circuits, can usually be considered as interconnections of lumped elements, such as sources, resistors, inductors, and capacitors. Sources are active electrical elements, which can provide energy to the circuit and serve as the inputs. Resistors, inductors, and capacitors can store or dissipate energy available in the circuit; however, they cannot produce energy. They are referred to as passive electrical elements.

Two primary variables used to describe the dynamic behavior of an electrical circuit are current and voltage. Current is the time rate of change of charge,

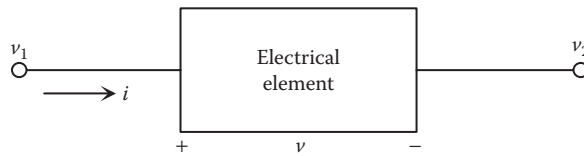
$$i = \frac{dq}{dt} \quad (6.1)$$

where:

q is charge in coulomb (C)

i is current in ampere (A)

For a two-terminal electrical element, the current entering one end of the element must be equal to the current leaving the other end. As shown in [Figure 6.1](#), an arrow denotes the direction in which the positive current (or charge) flows.

**FIGURE 6.1**

A two-terminal electrical element.

The voltage at a point in a circuit is a measure of the electrical potential difference between that point and a reference point called the ground. The unit of voltage is volt (V). If a point has the same electrical potential as the ground, it has a voltage of zero. For a two-terminal electrical element, the voltages at both ends are different. As shown in Figure 6.1, v_1 and v_2 denote the terminal voltages with respect to the ground, and

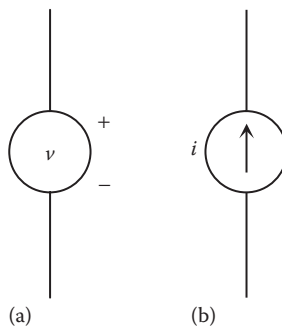
$$v = v_1 - v_2 \quad (6.2)$$

where v is the voltage across the element. The sense of the voltage v is indicated by plus and minus signs. The terminal with the plus sign has a higher voltage than that with the minus sign. The positive sense of the current associated with an electrical element is defined such that within the element, the positive current is assumed to flow from the high-voltage terminal to the low-voltage terminal. If the current flow has the same direction as the voltage drop, then the energy is supplied to the element. The supplied energy is either stored in the element or dissipated by the element. This type of electrical element is passive. For an active electrical element, current flows in the direction opposite to the voltage drop, and it supplies energy to the rest of the circuit. The power supplied to a passive element or generated by an active element is

$$P = vi \quad (6.3)$$

An active electrical element can be modeled as an ideal current source or an ideal voltage source, as shown in Figure 6.2. An ideal current source provides specified current no matter how much voltage is required by the circuit. An ideal voltage source provides specified voltage no matter how much current flows in the circuit.

To derive the dynamic model of an electrical circuit, it is important to understand the voltage–current relations for all passive electrical elements.

**FIGURE 6.2**

Active electrical elements: (a) ideal voltage source and (b) ideal current source.

6.1.1 Resistors

The voltage–current relation for a linear resistor, as shown in Figure 6.3, is an algebraic relationship,

$$v = Ri \quad (6.4)$$

where R is the resistance in units of ohm (Ω). Equation 6.4 is known as Ohm’s law, which states that the voltage and current of a linear resistor are directly proportional to each other. It is an empirical formula and can be obtained by a series of measurements.

A resistor dissipates energy by converting it into heat. The power dissipated by a linear resistor is given by

$$P = Ri^2 = \frac{v^2}{R} \quad (6.5)$$

In many electrical circuits, multiple resistors are used. They are arranged in different ways, such as series connections, parallel connections, or both. The equivalent resistance for several resistors arranged in any of these configurations can be obtained to simplify the modeling procedure.

Figure 6.4 shows a circuit with two resistors in series. It is known that the current (or charge) remains unchanged when crossing an electrical element. Thus, for the series connection, the current is the same through each resistor. Ohm’s law gives $v_1 = R_1i$ and $v_2 = R_2i$, where the voltages v_1 and v_2 also represent a measure of the energy required to move a charge through resistor R_1 or resistor R_2 , respectively. The total voltage required to move a charge across the two resistors is $v = v_1 + v_2 = (R_1 + R_2)i$. Comparing this result with the equivalent circuit, where $v = R_{eq}i$, we have

$$R_{eq} = R_1 + R_2 \quad (6.6)$$

A circuit with parallel resistors is shown in Figure 6.5. Note that the parallel resistors share the same terminals. Thus, the voltage across each resistor must be the same, $v_1 = v_2$. Ohm’s law gives $v = R_1i_1$ and $v = R_2i_2$, where i_1 and i_2 are currents through resistors R_1 and R_2 , respectively. Because of the conservation of charge, $i = i_1 + i_2 = v(1/R_1 + 1/R_2)$. Comparing this result with the equivalent circuit, where $i = v/R_{eq}$, we have

$$\frac{1}{R_{eq}} = \frac{1}{R_1} + \frac{1}{R_2} \quad (6.7)$$

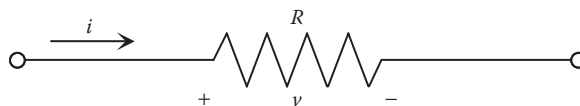
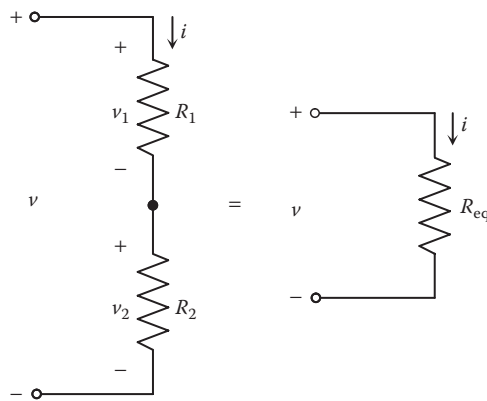
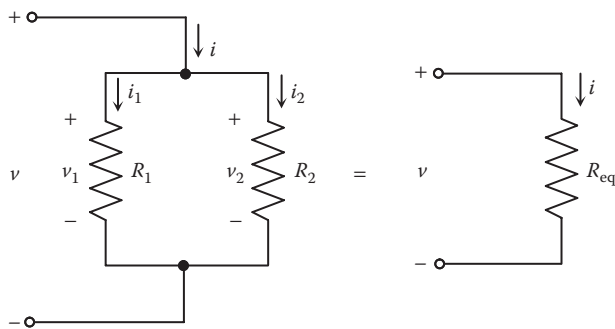


FIGURE 6.3

A resistor and its variables.

**FIGURE 6.4**

Equivalence for two resistors in series.

**FIGURE 6.5**

Equivalence for two resistors in parallel.

or

$$R_{\text{eq}} = \frac{R_1 R_2}{R_1 + R_2} \quad (6.8)$$

The results for series or parallel resistance can extend to n resistors. For a circuit of n resistors in series, the equivalent resistance is equal to the sum of all the individual resistances R_i :

$$R_{\text{eq}} = R_1 + R_2 + \cdots + R_n \quad (6.9)$$

For a system of n resistors in parallel, the reciprocal of the equivalent resistance R_{eq} is equal to the sum of all the reciprocals of the individual resistances R_i :

$$\frac{1}{R_{\text{eq}}} = \frac{1}{R_1} + \frac{1}{R_2} + \cdots + \frac{1}{R_n} \quad (6.10)$$

6.1.2 Inductors

Figure 6.6 shows the symbol for an inductor. The voltage–current relation for a linear inductor is

$$v = L \frac{di}{dt} \quad (6.11)$$

or

$$i = \frac{1}{L} \int v dt \quad (6.12)$$

where L is the inductance, and the unit is henry (H).

The energy supplied to an inductor is stored in its magnetic field. The stored energy can be derived by performing integration of the power

$$P = vi = \left(L \frac{di}{dt} \right) i \quad (6.13)$$

as follows:

$$E(t) = \int_0^t P(t) dt = \int_0^t Li(t) di = \frac{1}{2} Li^2 \Big|_0^t = \frac{1}{2} Li^2(t) - \frac{1}{2} Li^2(0) = \frac{1}{2} Li^2(t) \quad (6.14)$$

In Equation 6.14, it is assumed that the current through the inductor is zero at $t = 0$. Note that the energy stored in an inductor is dependent on the square of the current through the inductor and is independent of how the current is established.

6.1.3 Capacitors

Figure 6.7 shows the symbol for a capacitor, where C is the capacitance in units of farad (F). The capacitance is a measure of how much charge can be stored for a given voltage difference across the capacitor, and the mathematical description is $q = Cv$ or $v = q/C$. Note

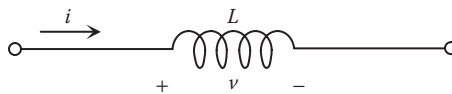


FIGURE 6.6

An inductor and its variables.

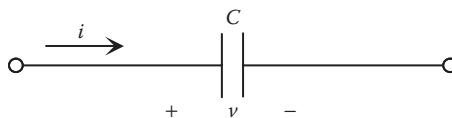


FIGURE 6.7

A capacitor and its variables.

that the charge q is related to the current via $i = dq/dt$ or $q = \int i dt$. Thus, the voltage–current relation for a capacitor is expressed as

$$v = \frac{1}{C} \int i dt \quad (6.15)$$

or

$$i = C \frac{dv}{dt} \quad (6.16)$$

A capacitor is also designed to store energy. The energy supplied to the capacitor is stored in its electrical field and can be derived by performing integration of the power

$$P = vi = v \left(C \frac{dv}{dt} \right) \quad (6.17)$$

as follows:

$$E(t) = \int_0^t P(t) dt = \int_0^t Cv(t)dv = \frac{1}{2} Cv^2 \Big|_0^t = \frac{1}{2} Cv^2(t) - \frac{1}{2} Cv^2(0) = \frac{1}{2} Cv^2(t) \quad (6.18)$$

In Equation 6.18, it is assumed that the voltage across the capacitor is zero at $t = 0$. Note that the energy stored in a capacitor is dependent on the square of the voltage across the capacitor and is independent of how the voltage is acquired.

PROBLEM SET 6.1

1. Determine the equivalent resistance R_{eq} for the circuit shown in [Figure 6.8](#).
2. Determine the equivalent resistance R_{eq} for the circuit shown in [Figure 6.9](#).
3. Determine the equivalent resistance R_{eq} for the circuit shown in [Figure 6.10](#). Assume that all resistors have the same resistance of R .
4. Determine the equivalent resistance R_{eq} for the circuit shown in [Figure 6.11](#). Assume that all resistors have the same resistance of R .

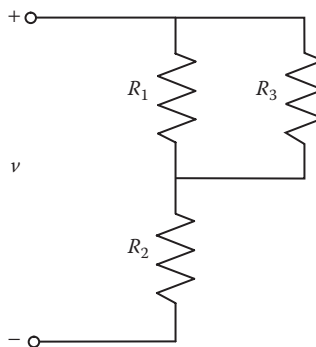


FIGURE 6.8
Problem 1.

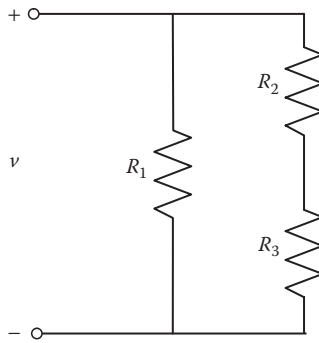


FIGURE 6.9
Problem 2.

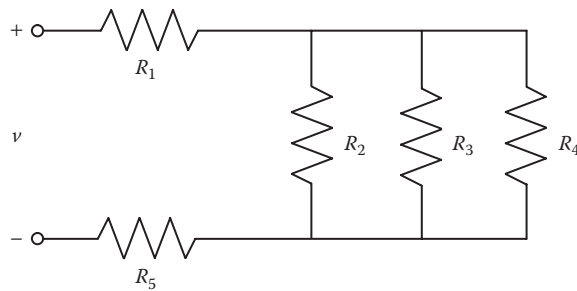


FIGURE 6.10
Problem 3.

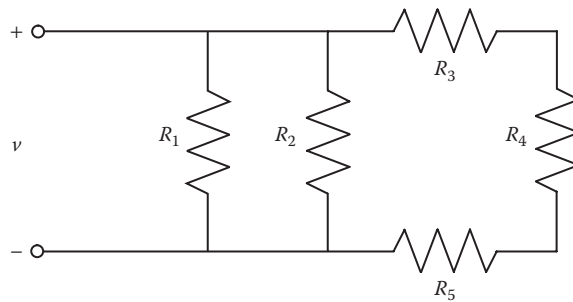
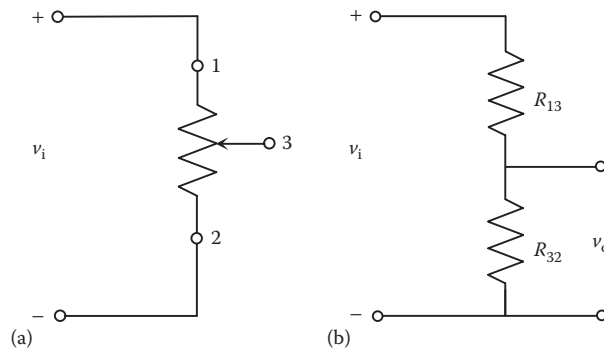
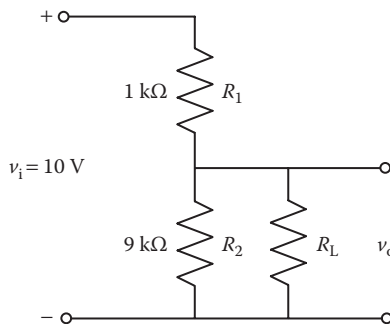


FIGURE 6.11
Problem 4.

5. A potentiometer is a variable resistor with three terminals. [Figure 6.12a](#) shows a potentiometer connected to a voltage source. The two end terminals are labeled as 1 and 2, and the adjustable terminal is labeled as 3. The potentiometer acts as a voltage divider, and the total resistance is separated into two parts, as shown in [Figure 6.12b](#), where R_{13} is the resistance between terminal 1 and terminal 3, and R_{32} is the resistance between terminal 3 and terminal 2. Determine the relationship between the input voltage v_i and the output voltage v_o .

**FIGURE 6.12**

Problem 5: (a) Potentiometer and (b) voltage divider.

**FIGURE 6.13**

Problem 6.

6. The output voltage of a voltage divider is not fixed; it varies according to the load.
 - a. Find the output voltage v_o in [Figure 6.13](#) for two different values of load resistance: (1) $R_L = 10 \text{ k}\Omega$ and (2) $R_L = 100 \text{ k}\Omega$.
 - b. Is it possible to get an output voltage v_o of greater than 9 V? If yes, determine the minimum value of the load resistance.
7. Consider an inductive divider shown in [Figure 6.14](#). For an alternating current (AC) input v_i , prove that the output voltage of the inductive voltage divider is
$$v_o = \frac{L_2}{L_1 + L_2} v_i.$$
8. Consider a capacitive divider shown in [Figure 6.15](#). For an AC input v_i , prove that the output voltage of the capacitive voltage divider is
$$v_o = \frac{C_1}{C_1 + C_2} v_i.$$
9. Consider a circuit of two inductors, L_1 and L_2 , in series. Prove that the equivalent inductance of the circuit is $L_{\text{eq}} = L_1 + L_2$.
10. Consider a circuit of two inductors, L_1 and L_2 , in parallel. Prove that the equivalent inductance of the circuit is
$$\frac{1}{L_{\text{eq}}} = \frac{1}{L_1} + \frac{1}{L_2}.$$
11. Consider a circuit of two capacitors, C_1 and C_2 , in series. Prove that the equivalent capacitance of the circuit is
$$\frac{1}{C_{\text{eq}}} = \frac{1}{C_1} + \frac{1}{C_2}.$$

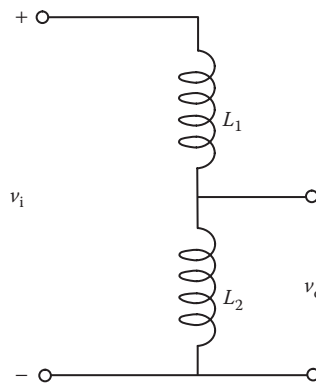


FIGURE 6.14
Problem 7.

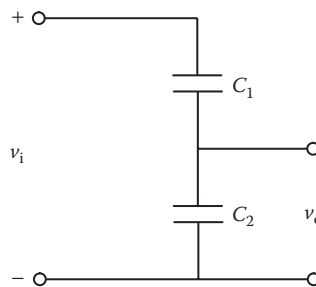


FIGURE 6.15
Problem 8.

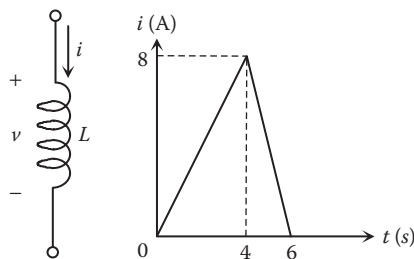


FIGURE 6.16
Problem 13.

12. Consider a circuit of two capacitors, C_1 and C_2 , in parallel. Prove that the equivalent capacitance of the circuit is $C_{eq} = C_1 + C_2$.
13. The current through an inductor of 5 mH is shown in [Figure 6.16](#). Find the voltage across the inductor. What is the energy stored in the inductor when (1) $t = 4$ s and (2) $t = 6$ s?
14. The voltage across a capacitor of 500 μ F is shown in [Figure 6.17](#). Find the current through the capacitor. What is the energy stored in the capacitor when (1) $t = 1$ s and (2) $t = 4$ s?

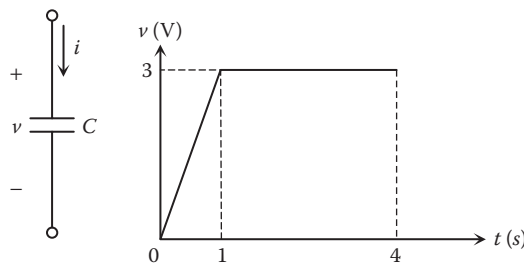


FIGURE 6.17
Problem 14.

6.2 Electric Circuits

When electrical elements are interconnected to form an electrical circuit, the dynamics model of the circuit can be developed using the voltage–current relations for electrical elements, along with two main physical laws. The two laws are known as Kirchhoff’s voltage law and Kirchhoff’s current law.

6.2.1 Kirchhoff’s Voltage Law

For a closed path, or a loop, in a circuit, Kirchhoff’s voltage law states that the algebraic sum of voltages around the loop must be zero,

$$\sum_j v_j = 0 \quad (6.19)$$

where v_j is the voltage across the j th element in the loop.

As an example, let us consider a series RLC circuit shown in [Figure 6.18](#), where a resistor (R), an inductor (L), and a capacitor (C) are connected in series. An ideal voltage

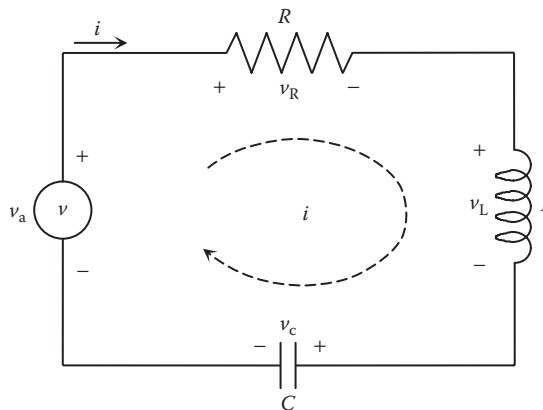


FIGURE 6.18
A series RLC circuit.

source provides the desired voltage to the circuit. The current flows from the high-voltage terminal of the source, crosses three passive elements, and enters the low-voltage terminal of the source. If we sum the voltages around the loop in a clockwise direction, a voltage drop occurs for each of the three passive elements, and a voltage gain occurs for the source. Assigning a positive sign to a voltage drop, and a negative sign to a voltage gain, Kirchhoff's voltage law gives $v_R + v_L + v_C - v_a = 0$ or $v_a = v_R + v_L + v_C$, which implies that the voltage of the source must equal the sum of the voltages across the resistor, the inductor, and the capacitor.

Example 6.1: A Series RLC Circuit

Consider the series RLC circuit shown in [Figure 6.18](#).

- Derive the differential equation in terms of the loop current i .
- Determine the transfer function $I(s)/V_a(s)$, which relates the source voltage $v_a(t)$ to the loop current $i(t)$. Assume that all the initial conditions are zero.
- Determine the transfer function $V_C(s)/V_a(s)$, which relates the source voltage $v_a(t)$ to the capacitor voltage $v_C(t)$. Assume that all the initial conditions are zero.

Solution

- Applying Kirchhoff's voltage law to the single loop along the clockwise direction gives

$$v_R + v_L + v_C - v_a = 0$$

For the series loop, the same current flows through each element. The expressions for v_R , v_L , and v_C are

$$v_R = Ri$$

$$v_L = L \frac{di}{dt}$$

$$v_C = \frac{1}{C} \int i dt$$

We then have

$$Ri + L \frac{di}{dt} + \frac{1}{C} \int i dt = v_a$$

Note that the above-mentioned equation is an integral-differential equation, not a differential equation. To eliminate the integral term, we take the time derivative of both sides of the equation. Rearranging the terms results in

$$L \frac{d^2i}{dt^2} + R \frac{di}{dt} + \frac{1}{C} i = \frac{dv_a}{dt}$$

which is a second-order differential equation for current i , with the time derivative of the applied voltage as the forcing function.

- Taking the Laplace transform of the previous differential equation yields

$$Ls^2I(s) + RsI(s) + \frac{1}{C}I(s) = sV_a(s)$$

The transfer function relating the input voltage $v_a(t)$ and the output current $i(t)$ is

$$\frac{I(s)}{V_a(s)} = \frac{s}{Ls^2 + Rs + (1/C)}$$

- c. Note that the capacitor voltage v_C does not appear explicitly in the differential equation. To determine the transfer function $V_C(s)/V_a(s)$, we use the result of the transfer function $I(s)/V_a(s)$ and apply the voltage–current relation for a capacitor

$$v_C = \frac{1}{C} \int i dt$$

which gives

$$V_C(s) = \frac{1}{Cs} I(s)$$

Thus, the transfer function relating the input voltage $v_a(t)$ and the capacitor voltage $v_C(t)$ is

$$\frac{V_C(s)}{V_a(s)} = \frac{1}{Cs} \frac{I(s)}{V_a(s)} = \frac{1}{LCs^2 + RCs + 1}$$

The circuit in [Figure 6.18](#) can also be modeled using a differential equation for charge q . Recall that current is the time rate of change of charge, $i = dq/dt$ or $q = \int i dt$. Rewriting the current-related terms in the equation $Ri + L(di/dt) + (1/C)\int i dt = v_a$ in terms of q gives

$$L \frac{d^2 q}{dt^2} + R \frac{dq}{dt} + \frac{1}{C} q = v_a$$

which is a second-order differential equation for the charge $q(t)$, with the applied voltage as the forcing function.

6.2.2 Kirchhoff's Current Law

When the terminals of two or more circuit elements are connected together, the common junction is referred to as a node. For a node in a circuit, Kirchhoff's current law states that the sum of the currents entering the node must be equal to the sum of the currents leaving that node. If we assign a positive sign to the current entering the node and a negative sign to the current leaving the node, the algebraic sum of the currents at the node must be zero,

$$\sum_j i_j = 0 \tag{6.20}$$

where i_j is the current of the j th element at the node.

Example 6.2: A Parallel RLC Circuit

Consider the parallel RLC circuit shown in [Figure 6.19](#), in which an ideal current source supplies the desired current to the circuit.

- Derive the differential equation relating the input current i_a to the output voltage v_o .
- Determine the transfer function $V_o(s)/I_a(s)$, which relates the input current $i_a(t)$ to the output voltage $v_o(t)$. Assume that all the initial conditions are zero.
- Determine the transfer function $I_L(s)/I_a(s)$, which relates the input current $i_a(t)$ to the current through the inductor $i_L(t)$. Assume that all the initial conditions are zero.

Solution

- The currents, i_R , i_L , and i_C are defined in [Figure 6.19](#). Each passive element has one terminal connected to the ground and the other terminal connected to a common node. We can apply Kirchhoff's law to either the ground or node 1. Applying Kirchhoff's current law to node 1 gives

$$i_a - i_R - i_L - i_C = 0$$

For the parallel connection, the voltages across all three elements are the same. The expressions for i_R , i_L , and i_C are

$$i_R = \frac{v_o}{R}$$

$$i_L = \frac{1}{L} \int v_o dt$$

$$i_C = C \frac{dv_o}{dt}$$

We then have

$$\frac{v_o}{R} + \frac{1}{L} \int v_o dt + C \frac{dv_o}{dt} = i_a$$

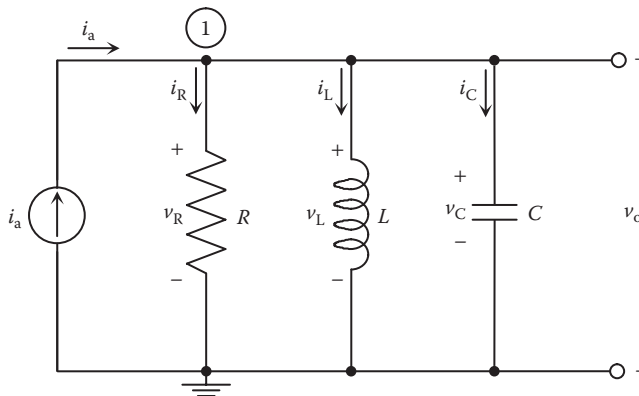


FIGURE 6.19

A parallel RLC circuit.

which is an integral–differential equation. To eliminate the integral term, we take the time derivative of both sides of the equation. Rearranging the terms results in

$$C \frac{d^2 v_o}{dt^2} + \frac{1}{R} \frac{dv_o}{dt} + \frac{1}{L} v_o = \frac{di_a}{dt}$$

which is a second-order differential equation for the output voltage $v_o(t)$.

b. Taking the Laplace transform of the previous differential equation yields

$$Cs^2 V_o(s) + \frac{1}{R} s V_o(s) + \frac{1}{L} V_o(s) = s I_a(s)$$

The transfer function relating the input current $i_a(t)$ and the output voltage $v_o(t)$ is

$$\frac{V_o(s)}{I_a(s)} = \frac{s}{Cs^2 + (1/R)s + (1/L)}$$

c. To find the transfer function $I_L(s)/I_a(s)$, note that

$$i_L = \frac{1}{L} \int v_o dt$$

which gives

$$I_L(s) = \frac{1}{Ls} V_o(s)$$

Thus,

$$\frac{I_L(s)}{I_a(s)} = \frac{1}{Ls} \frac{V_o(s)}{I_a(s)} = \frac{1}{LCs^2 + (L/R)s + 1}$$

The above-mentioned two simple examples illustrate how one can develop a differential equation for a series or parallel RLC circuit by using Kirchhoff's voltage law or Kirchhoff's current law. However, if circuit components are series connected in some parts and parallel in others, we must selectively apply Kirchhoff's voltage law and Kirchhoff's current law to obtain the desired differential equation.

Example 6.3: An RC Circuit

Consider the circuit shown in [Figure 6.20](#). Derive the differential equation relating the output voltage $v_o(t)$ to the input voltage $v_a(t)$.

Solution

The currents through the three passive elements (i_{R_1} , i_{R_2} , and i_C) are defined in [Figure 6.20](#). Applying Kirchhoff's current law gives

$$i_{R_1} = i_{R_2} + i_C$$

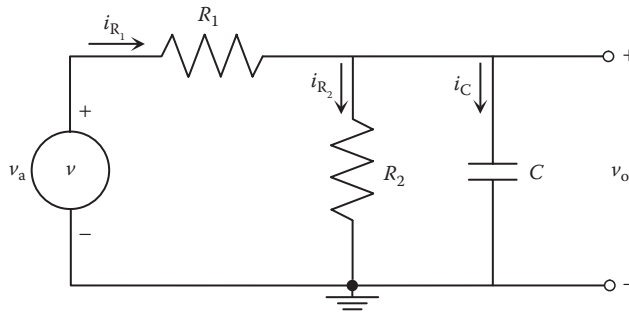


FIGURE 6.20
An RC circuit.

Applying voltage–current relations and Kirchhoff’s voltage law give

$$i_{R_1} = \frac{v_{R_1}}{R_1} = \frac{v_a - v_o}{R_1}$$

$$i_{R_2} = \frac{v_{R_2}}{R_2} = \frac{v_o}{R_2}$$

$$i_C = C \frac{dv_C}{dt} = C \frac{dv_o}{dt}$$

Substituting the above-mentioned expressions into the Kirchhoff’s current law, we have

$$\frac{v_a - v_o}{R_1} = \frac{v_o}{R_2} + C \frac{dv_o}{dt}$$

Rearranging the terms results in

$$C \frac{dv_o}{dt} + \left(\frac{1}{R_1} + \frac{1}{R_2} \right) v_o = \frac{1}{R_1} v_a$$

As shown in Example 6.3, the procedure for deriving the differential equation of a circuit consists of applying Kirchhoff’s laws and voltage–current relations for the components of the circuit. However, obtaining a set of equations for more complicated circuits is not that easy. Instead, a formal method is needed that produces a small, simple set of equations leading directly to the input–output relation. The two commonly used methods are the node method (which relies on Kirchhoff’s current law) and the loop method (which relies on Kirchhoff’s voltage law).

6.2.3 Node Method

If a node in a circuit is chosen as the reference, any other node can be assigned a voltage, which is defined between this node and the reference. This common reference node is usually referred to as the ground. To apply the node method to a circuit, we start by labeling all currents at each node whose voltage is unknown. The current

through each passive circuit element is expressed in terms of the node voltages, using the voltage–current relations given in [Section 6.1](#). We then apply Kirchhoff's current law to each node, and the resulting set of equations can be combined to obtain the complete model of the circuit. The following two examples show details of the node-voltage method.

Example 6.4: Circuit Modeling Using the Node Method: One Node

Consider the circuit shown in [Figure 6.21](#). Derive the differential equation relating the output voltage $v_o(t)$ to the input voltage $v_a(t)$.

Solution

Note that the voltage at node 1 is unknown and we denote it as v_1 . All currents entering or leaving node 1 are labeled as shown in [Figure 6.21](#). Applying Kirchhoff's current law to node 1 gives

$$i_L - i_R - i_C = 0$$

Expressing the current through each element in terms of the node voltage, we have

$$\frac{1}{L} \int (v_a - v_1) dt - \frac{v_1 - 0}{R} - C \frac{d}{dt} (v_1 - 0) = 0$$

Differentiating the above-mentioned equation with respect to time results in

$$\frac{v_a - v_1}{L} - \frac{\dot{v}_1}{R} - C \ddot{v}_1 = 0$$

Because the node voltage v_1 is essentially the output voltage v_o , the aforementioned equation can be rewritten as

$$RLC\ddot{v}_o + L\dot{v}_o + Rv_o = Rv_a$$

which is the input–output equation relating the applied voltage v_a and the output voltage v_o .

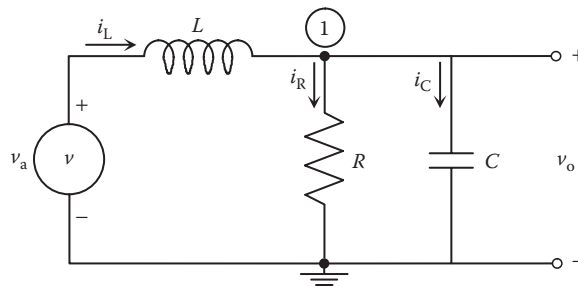


FIGURE 6.21
A circuit containing one node.

Example 6.5: Circuit Modeling Using the Node Method: Two Nodes

Consider the circuit shown in Figure 6.22.

- Using the node method, derive the differential equations of the system in terms of node voltages.
- Determine the transfer function $V_o(s)/V_a(s)$. Assume that all initial conditions are zero.

Solution

- Note that the voltages at node 1 and node 2 are unknown. Denote the voltage at node 1 as v_1 and the voltage at node 2 as v_2 . All currents entering or leaving node 1 and node 2 are labeled as shown in Figure 6.22.

At node 1, applying Kirchhoff's current law gives

$$i_{L_1} - i_{L_2} - i_C = 0$$

Expressing the current through each element in terms of the node voltages, we have

$$\frac{1}{L_1} \int (v_a - v_1) dt - \frac{1}{L_2} \int (v_1 - v_2) dt - C \frac{dv_1}{dt} = 0$$

Differentiating the previous equation with respect to time results in

$$\frac{1}{L_1} (v_a - v_1) - \frac{1}{L_2} (v_1 - v_2) - C \ddot{v}_1 = 0$$

which can be rearranged to give the first differential equation

$$C \ddot{v}_1 + \left(\frac{1}{L_1} + \frac{1}{L_2} \right) v_1 - \frac{1}{L_2} v_2 = \frac{1}{L_1} v_a$$

Similarly, applying Kirchhoff's current law to node 2 yields

$$i_{L_2} - i_R = 0$$

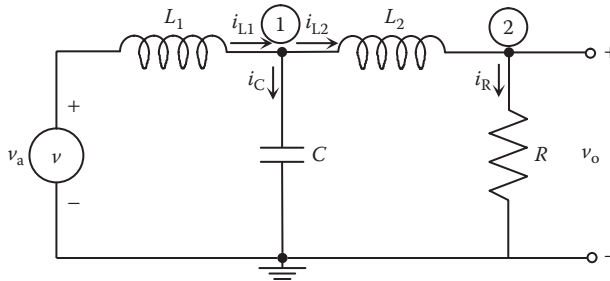


FIGURE 6.22

A circuit containing two nodes.

Expressing the current through each element in terms of the node voltages, we have

$$\frac{1}{L_2} \int (v_1 - v_2) dt - \frac{v_2}{R} = 0$$

Differentiating the previous equation with respect to time results in

$$\frac{1}{L_2} (v_1 - v_2) - \frac{1}{R} \dot{v}_2 = 0$$

which can be rearranged to give the second differential equation

$$\frac{1}{R} \dot{v}_2 - \frac{1}{L_2} v_1 + \frac{1}{L_2} v_2 = 0$$

In second-order matrix form, we have

$$\begin{bmatrix} C & 0 \\ 0 & 0 \end{bmatrix} \begin{Bmatrix} \ddot{v}_1 \\ \ddot{v}_2 \end{Bmatrix} + \begin{bmatrix} 0 & 0 \\ 0 & \frac{1}{R} \end{bmatrix} \begin{Bmatrix} \dot{v}_1 \\ \dot{v}_2 \end{Bmatrix} + \begin{bmatrix} \frac{1}{L_1} + \frac{1}{L_2} & -\frac{1}{L_2} \\ -\frac{1}{L_2} & \frac{1}{L_2} \end{bmatrix} \begin{Bmatrix} v_1 \\ v_2 \end{Bmatrix} = \begin{Bmatrix} \frac{1}{L_1} \\ 0 \end{Bmatrix} v_a$$

b. Taking Laplace transform of the above-mentioned two equations gives

$$\begin{bmatrix} Cs^2 + \left(\frac{1}{L_1} + \frac{1}{L_2} \right) & -\frac{1}{L_2} \\ -\frac{1}{L_2} & \frac{1}{R}s + \frac{1}{L_2} \end{bmatrix} \begin{Bmatrix} V_1(s) \\ V_2(s) \end{Bmatrix} = \begin{Bmatrix} \frac{1}{L_1} \\ 0 \end{Bmatrix} V_a(s)$$

Note that $v_o = v_2$. Applying Cramer's rule gives

$$\frac{V_o(s)}{V_a(s)} = \frac{V_2(s)}{V_a(s)} = \frac{R}{L_1 L_2 C s^3 + R L_1 C s^2 + (L_1 + L_2) s + R}$$

6.2.4 Loop Method

In a circuit carrying some current, there exists at least one loop. Starting from any loop, a current circulating that loop can be assigned. For any additional loop containing at least one new element that is not in any previous loops, a new loop current can be assigned. To apply the loop method to a circuit, we start by assigning each loop current. Generally, assume that all unknown currents flow in the clockwise direction, and all known currents follow the directions of the current sources. The voltage across each passive element is expressed in terms of loop currents, using the voltage-current relations given in [Section 6.1](#). We then apply Kirchhoff's voltage law to each loop with unknown current, and the resulting set of equations can be combined to obtain the complete model of the circuit. The following example shows the details of the loop-current method.

Example 6.6: Circuit Modeling Using the Loop Method: Two Loops

Reconsider the circuit in Figure 6.22 and solve Example 6.5 by using the loop method.

Solution

Assign loop currents as shown in Figure 6.23. Note that there are two loops with unknown currents.

For loop 1, applying Kirchhoff's voltage law gives

$$v_{L_1} + v_C - v_a = 0$$

Expressing the voltage across each element in terms of the loop currents, we have

$$L_1 \frac{di_1}{dt} + \frac{1}{C} \int (i_1 - i_2) dt - v_a = 0$$

Differentiating the above-mentioned equation with respect to time results in

$$L_1 \frac{d^2 i_1}{dt^2} + \frac{1}{C} i_1 - \frac{1}{C} i_2 = \frac{dv_a}{dt}$$

Similarly, applying Kirchhoff's voltage law to loop 2 gives

$$v_{L_2} + v_R + v_C = 0$$

$$L_2 \frac{di_2}{dt} + Ri_2 + \frac{1}{C} \int (i_2 - i_1) dt = 0$$

Differentiating the previous equation with respect to time results in

$$L_2 \frac{d^2 i_2}{dt^2} + R \frac{di_2}{dt} - \frac{1}{C} i_1 + \frac{1}{C} i_2 = 0$$

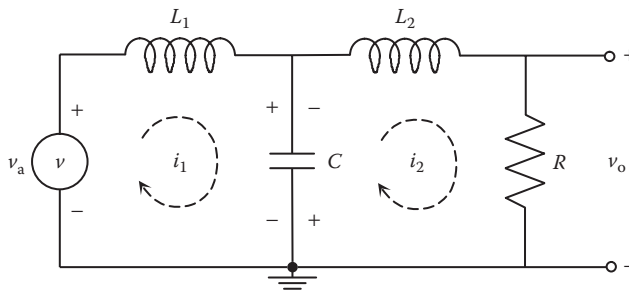


FIGURE 6.23

A circuit containing two loops.

The above-mentioned two differential equations can be written in second-order matrix form, as

$$\begin{bmatrix} L_1 & 0 \\ 0 & L_2 \end{bmatrix} \begin{Bmatrix} \frac{d^2 i_1}{dt^2} \\ \frac{d^2 i_2}{dt^2} \end{Bmatrix} + \begin{bmatrix} 0 & 0 \\ 0 & R \end{bmatrix} \begin{Bmatrix} \frac{di_1}{dt} \\ \frac{di_2}{dt} \end{Bmatrix} + \begin{bmatrix} \frac{1}{C} & -\frac{1}{C} \\ -\frac{1}{C} & \frac{1}{C} \end{bmatrix} \begin{Bmatrix} i_1 \\ i_2 \end{Bmatrix} = \begin{Bmatrix} \frac{dv_a}{dt} \\ 0 \end{Bmatrix}$$

Note that the output voltage v_o is related to the current through the resistor, that is, $v_o = Ri_2$ or $V_o(s) = RI_2(s)$. Taking the Laplace transform of the differential equations gives

$$\begin{bmatrix} L_1 s^2 + \frac{1}{C} & -\frac{1}{C} \\ -\frac{1}{C} & L_2 s^2 + Rs + \frac{1}{C} \end{bmatrix} \begin{Bmatrix} I_1(s) \\ I_2(s) \end{Bmatrix} = \begin{Bmatrix} sV_a(s) \\ 0 \end{Bmatrix}$$

Using Cramer's rule, we have

$$I_2(s) = \frac{1}{L_1 L_2 C s^3 + R L_1 C s^2 + (L_1 + L_2)s + R}$$

Thus, the transfer function $V_o(s)/V_a(s) = RI_2(s)/V_a(s)$, gives the same result obtained previously in Example 6.5.

Example 6.6 shows that the loop method is similar to the node method. The choice between the two methods is often made based on the circuit at hand. For example, considering the circuit in [Figure 6.21](#), there is one independent node, but two independent loops. Therefore, the node method is expected to be easier to apply (see Example 6.4). The reader can try the loop method to solve the same problem. We will emphasize the node method to obtain mathematical models of circuits in this book.

6.2.5 State Variables of Circuits

To represent a circuit model in state-space form, we need to choose an appropriate set of state variables, whose time derivatives are expressed in terms of the state variables and inputs. Since the choice of state variables is not unique, it is difficult to identify the appropriate states for expressing a circuit in state-space form. Here, we introduce a customary choice of state variables by identifying the energy storage elements. As stated in [Section 6.1](#), both inductors and capacitors can store energy. In a given circuit, knowledge of the voltage signals across capacitors and of the current signals through inductors is sufficient to calculate other circuit variables by using only algebraic equations. Generally, inductor currents and capacitor voltages are continuous in nature and are often chosen as the state variables. To determine the state-space form of an electric circuit, we need to find the expression of di_L/dt or dv_C/dt for each inductor or capacitor. Based on the voltage-current relations, we have

$$\frac{di_L}{dt} = \frac{v_L}{L} \quad (6.21)$$

and

$$\frac{dv_C}{dt} = \frac{i_C}{C} \quad (6.22)$$

where:

v_L is the inductor voltage

i_C is the capacitor current

Thus, the problem is converted to expressing v_L and i_C in terms of state variables and inputs, using Kirchhoff's laws and voltage–current relations for electrical elements.

Example 6.7: State-Variable Model of the Circuit in Example 6.4

Reconsider the circuit shown in [Figure 6.21](#).

- Derive the state-variable model with inductor currents and capacitor voltages as states. The input is the applied voltage v_a , and the output is the voltage across the capacitor C .
- Based on the state-space form obtained in Part (a), determine the differential equation relating the output voltage $v_o(t)$ to the input voltage $v_a(t)$.

Solution

- We first label the nodes and currents as we did in Example 6.4. Note that the circuit has two independent energy storage elements, L and C . This implies that two state variables are needed, and they are

$$x_1 = i_L, \quad x_2 = v_C$$

Their time derivatives are

$$\dot{x}_1 = \frac{di_L}{dt} = \frac{1}{L} v_L$$

$$\dot{x}_2 = \frac{dv_C}{dt} = \frac{1}{C} i_C$$

We need to express the voltage across the inductor, v_L , and the current through the capacitor, i_C , in terms of the state variables and the input. Note that

$$\begin{array}{c} \text{Kirchhoff's voltage law} \\ v_L = v_a - v_C = u - x_2 \end{array}$$

and at node 1

$$\begin{array}{c} \text{Kirchhoff's current law} \\ i_C = i_L - i_R = \frac{v_L}{R} = \frac{v_a - v_C}{R} = \frac{u - x_2}{R} \end{array}$$

Thus, the complete set of two state-variable equations is

$$\dot{x}_1 = \frac{1}{L} v_L = -\frac{1}{L} x_2 + \frac{1}{L} u$$

$$\dot{x}_2 = \frac{1}{C} i_C = \frac{1}{C} x_1 - \frac{1}{RC} x_2$$

The output equation is

$$y = v_o = v_C = x_2$$

The state equation and the output equation can be written in matrix form as follows:

$$\begin{aligned} \begin{Bmatrix} \dot{x}_1 \\ \dot{x}_2 \end{Bmatrix} &= \begin{bmatrix} 0 & -\frac{1}{L} \\ \frac{1}{C} & -\frac{1}{RC} \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix} + \begin{bmatrix} 1/L \\ 0 \end{bmatrix} u \\ y &= \begin{bmatrix} 0 & 1 \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix} + 0 \cdot u \end{aligned}$$

- b. Note that $Y(s) = V_o(s)$ and $U(s) = V_a(s)$. As presented in [Section 4.4](#), the state-space form can be converted to a transfer function by using

$$\frac{Y(s)}{U(s)} = \mathbf{C}(s\mathbf{I} - \mathbf{A})^{-1}\mathbf{B} + D$$

Substituting $\mathbf{A}$, $\mathbf{B}$, $\mathbf{C}$, and D matrices obtained in Part (a) gives

$$\frac{V_o(s)}{V_a(s)} = \begin{bmatrix} 0 & 1 \end{bmatrix} \begin{bmatrix} s & \frac{1}{L} \\ -\frac{1}{C} & s + \frac{1}{RC} \end{bmatrix}^{-1} \begin{bmatrix} \frac{1}{L} \\ 0 \end{bmatrix} = \frac{R}{RLCs^2 + Ls + R}$$

or

$$(RLCs^2 + Ls + R)V_o(s) = RV_a(s)$$

Taking the inverse Laplace transform gives the same input–output equation as the one obtained in Example 6.4.

As shown in the example above, if inductor currents and capacitor voltages are chosen as state variables, the key to derive the state-variable model of a circuit is to express all other circuit variables by using only algebraic equations by applying Kirchhoff's voltage law, Kirchhoff's current law, Ohm's law, and the properties of series and parallel connections.

Example 6.8: State-Variable Model of the Circuit in Example 6.5

Reconsider the circuit shown in [Figure 6.22](#). Derive the state-variable model with inductor currents and capacitor voltages as states. The input is the applied voltage $v_a(s)$, and the output is the voltage across the resistor R .

Solution

Refer to [Figure 6.22](#) shown in Example 6.5. Note that the circuit has three independent energy storage elements, L_1 , L_2 , and C . This implies that three states are needed, and they are

$$x_1 = i_{L_1}, \quad x_2 = i_{L_2}, \quad x_3 = v_C$$

Their time derivatives are

$$\dot{x}_1 = \frac{di_{L_1}}{dt} = \frac{1}{L_1} v_{L_1}$$

$$\dot{x}_2 = \frac{di_{L_2}}{dt} = \frac{1}{L_2} v_{L_2}$$

$$\dot{x}_3 = \frac{dv_C}{dt} = \frac{1}{C} i_C$$

where:

$$v_{L_1} = v_a - v_1 = v_a - v_C = u - x_3$$

$$v_{L_2} = v_1 - v_2 = v_C - Ri_{L_2} = x_3 - Rx_2$$

$$i_C = i_{L_1} - i_{L_2} = x_1 - x_2$$

Thus, the complete set of three state-variable equations is

$$\dot{x}_1 = -\frac{1}{L_1} x_3 + \frac{1}{L_1} u$$

$$\dot{x}_2 = -\frac{R}{L_2} x_2 + \frac{1}{L_2} x_3$$

$$\dot{x}_3 = \frac{1}{C} x_1 - \frac{1}{C} x_2$$

The output equation is

$$y = v_o = v_R = Ri_{L_2} = Rx_2$$

The state equation and the output equation can be written in matrix form as follows:

$$\begin{Bmatrix} \dot{x}_1 \\ \dot{x}_2 \\ \dot{x}_3 \end{Bmatrix} = \begin{bmatrix} 0 & 0 & -1/L_1 \\ 0 & -R/L_2 & 1/L_2 \\ 1/C & -1/C & 0 \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \\ x_3 \end{Bmatrix} + \begin{bmatrix} 1/L_1 \\ 0 \\ 0 \end{bmatrix} u$$

$$y = \begin{bmatrix} 0 & R & 0 \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \\ x_3 \end{Bmatrix} + 0 \cdot u$$

PROBLEM SET 6.2

1. Consider the first-order RC circuit shown in [Figure 6.24](#).
 - a. Derive the input-output differential equation relating v_C and v_a .
 - b. Determine the transfer function $I(s)/V_a(s)$, which relates the loop current $i(t)$ to the applied voltage $v_a(t)$. Assume that all the initial conditions are zero.

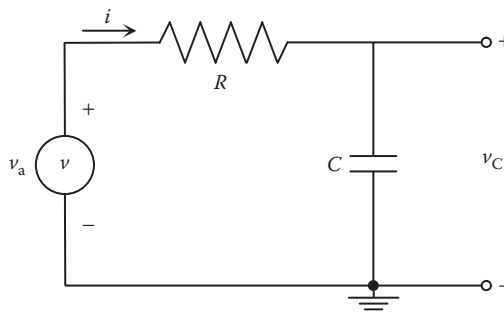


FIGURE 6.24
Problem 1.

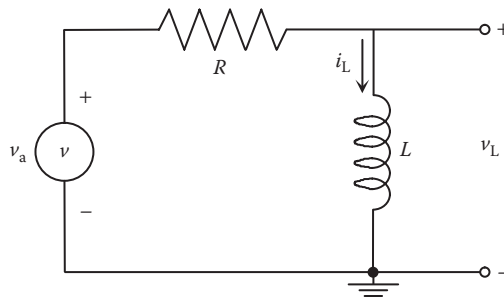


FIGURE 6.25
Problem 2.

2. Consider the first-order RL circuit shown in [Figure 6.25](#).
 - a. Derive the input–output differential equation relating i_L and v_a .
 - b. Determine the transfer function $V_L(s)/V_a(s)$, which relates the voltage across the inductor $v_L(t)$ to the applied voltage $v_a(t)$. Assume that all the initial conditions are zero.
3. Consider the circuit shown in [Figure 6.26](#). Use the node method to derive the input–output differential equation relating v_o and v_a .
4. Repeat Problem 3, using the loop method.

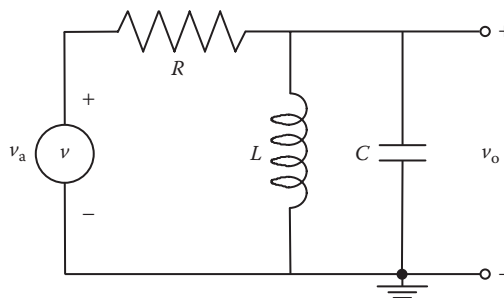


FIGURE 6.26
Problem 3.

5. Consider the circuit shown in Figure 6.27. Use the node method to derive the input–output differential equation relating i and v_a .
6. Repeat Problem 5, using the loop method.
7. Consider the circuit shown in Figure 6.28. Use the node method to derive the input–output differential equation relating v_o and v_a .
8. Repeat Problem 7, using the loop method.
9. Consider the circuit shown in Figure 6.29. Use the node method to derive the transfer function $V_o(s)/V_a(s)$. Assume that all initial conditions are zero.
10. Reconsider the circuit shown in Figure 6.29. Use the loop method to derive the transfer function $V_o(s)/V_a(s)$. Assume that all initial conditions are zero.
11. Consider the circuit shown in Figure 6.26. Determine a suitable set of state variables and obtain the state-space representation with v_o as the output.
12. Repeat Problem 11 for the circuit shown in Figure 6.27.
13. Repeat Problem 11 for the circuit shown in Figure 6.28.
14. Repeat Problem 11 for the circuit shown in Figure 6.29.
15. Repeat Problem 11 for the circuit shown in Figure 6.30.
16. Repeat Problem 11 for the circuit shown in Figure 6.31.

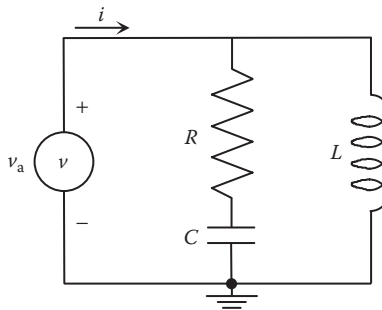


FIGURE 6.27
Problem 5.

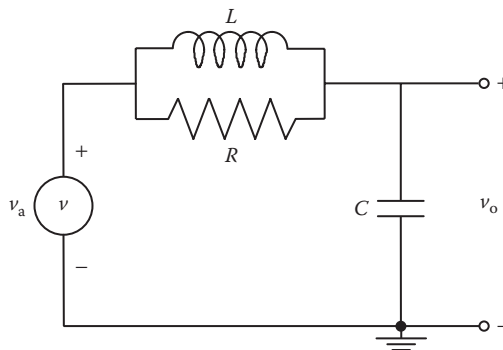


FIGURE 6.28
Problem 7.

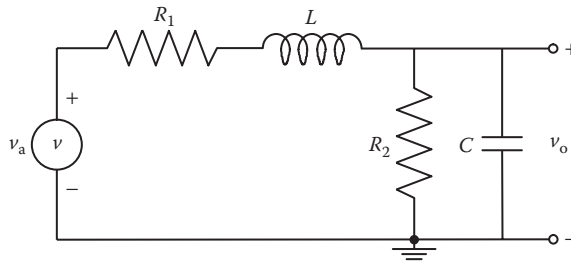


FIGURE 6.29
Problem 9.

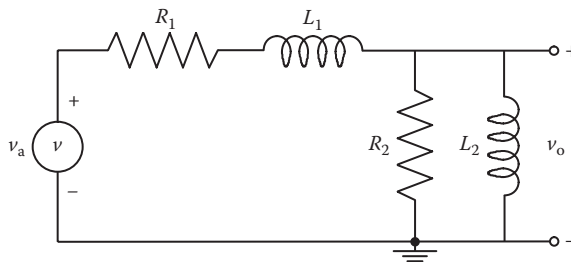


FIGURE 6.30
Problem 15.

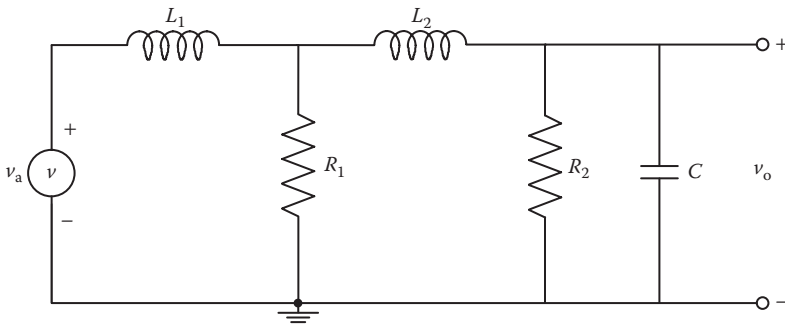
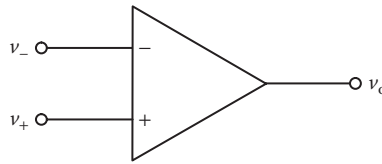


FIGURE 6.31
Problem 16.

6.3 Operational Amplifiers

An op-amp is an electronic element that is used to amplify electrical signals and drive physical devices. Figure 6.32 shows the schematic diagram of an op-amp, which is a voltage amplifier with a high gain K . Unlike the electrical elements discussed in earlier sections, op-amps have more than two terminals. The diagram in Figure 6.32 does not show all the terminals connected to the physical devices. It only shows a pair of input terminals and one output terminal. The output voltage is

$$v_o = K(v_+ - v_-) \quad (6.23)$$

**FIGURE 6.32**

An op-amp.

where K is a very large positive number, typically 10^5 – 10^6 . Because the output voltage v_o must be a finite number and K is very large, the voltage difference between the input terminals must approach zero. Thus,

$$v_+ \approx v_- \quad (6.24)$$

which is considered to be the op-amp equation.

Note that the diagram in Figure 6.32 is a simple symbol for an op-amp, which typically contains many resistors, inductors, and capacitors built on an integrated chip.

Example 6.9: An Op-Amp Multiplier

Consider the op-amp circuit shown in Figure 6.33, in which one resistor R_2 is in parallel connection with an op-amp, and the resulting parallel circuit is in series connection with another resistor R_1 . Determine the relation between the input voltage v_i and the output voltage v_o . Assume that the current drawn by the op-amp is very small.

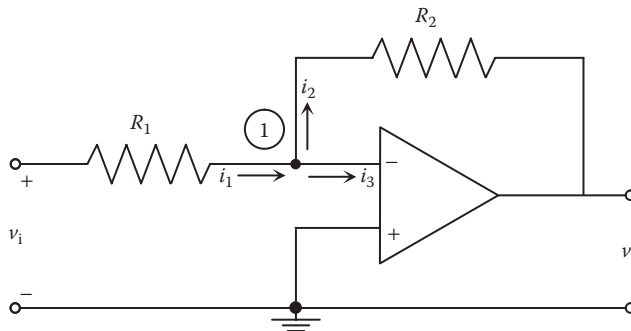
Solution

Label the currents at the nodes with unknown voltages. The system has only one significant node: node 1. Applying Kirchhoff's current law to node 1 gives

$$i_1 - i_2 - i_3 = 0$$

Because the current drawn by the op-amp is very small, that is, $i_3 \approx 0$, we have

$$i_1 \approx i_2$$

**FIGURE 6.33**

An op-amp multiplier circuit.

Using the voltage–current relation for each resistor yields

$$\frac{v_i - v_1}{R_1} = \frac{v_1 - v_o}{R_2}$$

Note that the input terminal marked with the plus sign is connected to the ground. From the op-amp equation $v_+ \approx v_-$, we have

$$v_1 = v_- \approx v_+ = 0$$

Thus, the relation between the input voltage v_i and the output voltage v_o is

$$\frac{v_i}{R_1} = -\frac{v_o}{R_2}$$

or

$$v_o = -\frac{R_2}{R_1} v_i$$

This circuit is known as an op-amp multiplier and is widely used in control systems. Op-amps can also be used for integrating and differentiating signals.

Example 6.10: An Op-Amp Differentiator

Consider the op-amp circuit shown in [Figure 6.34](#). Derive the differential equation relating the input voltage v_i and the output voltage v_o .

Solution

Note that the current drawn by the op-amp is very small. Applying Kirchhoff's current law to node 1 gives

$$i_1 = i_2$$

$$C \frac{d}{dt}(v_i - v_1) = \frac{v_1 - v_o}{R}$$

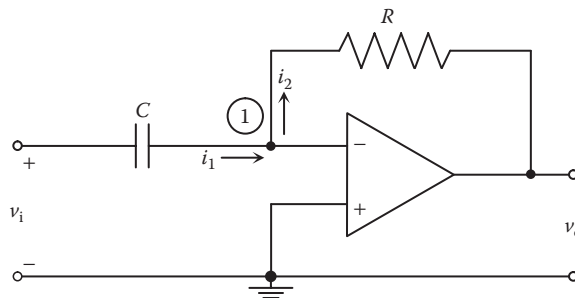


FIGURE 6.34
An op-amp differentiator circuit.

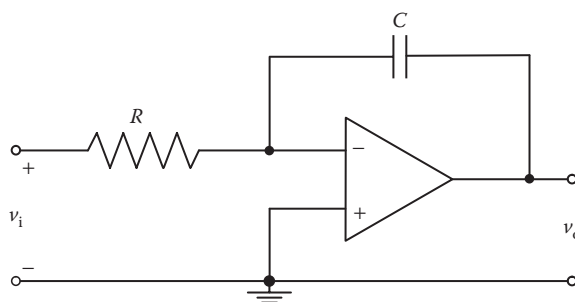


FIGURE 6.35
An op-amp integrator circuit.

Because the input terminal marked with the plus sign is connected to the ground, the op-amp equation yields $v_i = v_- \approx v_+ = 0$. Thus, the differential equation for the op-amp circuit in Figure 6.34 is

$$C\dot{v}_i = -\frac{v_o}{R}$$

or

$$v_o = -RC\dot{v}_i$$

This implies that the output voltage v_o is proportional to the time derivative of the input voltage v_i . The circuit in Figure 6.34 is therefore called a differentiator. Switching the resistor and the capacitor in Figure 6.34 results in an op-amp integrator shown in Figure 6.35. The relation between the output voltage and the input voltage is $\dot{v}_o = -v_i/(RC)$. We leave the derivation as an exercise for the reader.

Example 6.11: An Op-Amp Circuit

Consider the op-amp circuit shown in Figure 6.36. Derive the differential equation relating the output voltage $v_o(t)$ and the input voltage $v_i(t)$.

Solution

Note that the current flowing into the input terminal of the op-amp is very small. Applying Kirchhoff's current law to node 1 yields

$$i_{R_1} + i_{C_1} - i_{R_2} - i_{C_2} = 0$$

Using the voltage–current relations for electrical elements to express each term in the equation, we obtain

$$\frac{v_i - v_1}{R_1} + C_1 \frac{d}{dt}(v_i - v_1) - \frac{v_1 - v_o}{R_2} - C_2 \frac{d}{dt}(v_1 - v_o) = 0$$

Because $v_+ = 0$, the op-amp equation yields $v_i = v_- \approx v_+ = 0$. Substituting this into the previous equation results in

$$\frac{v_i}{R_1} + C_1 \dot{v}_i - \frac{-v_o}{R_2} - C_2 (-\dot{v}_o) = 0$$

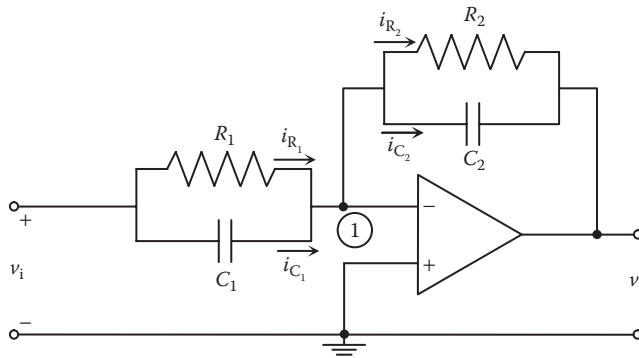


FIGURE 6.36
An op-amp circuit.

which can be rearranged into

$$C_2 \dot{v}_o + \frac{1}{R_2} v_o = -C_1 \dot{v}_i - \frac{1}{R_1} v_i$$

PROBLEM SET 6.3

1. The op-amp circuit shown in [Figure 6.37](#) is a summing amplifier. Determine the relation between the input voltages v_1 , v_2 , and the output voltage v_o .
2. The op-amp circuit shown in [Figure 6.38](#) is a difference amplifier. Determine the relation between the input voltages v_1 , v_2 , and the output voltage v_o .
3. The op-amp circuit shown in [Figure 6.39](#) is a noninverting amplifier. Determine the relation between the input voltage v_i and the output voltage v_o .
4. Consider the op-amp integrator circuit shown in [Figure 6.35](#). Derive the differential equation relating the input voltage v_i and the output voltage v_o .
5. Consider the op-amp circuit shown in [Figure 6.40](#). Derive the differential equation relating the input voltage v_i and the output voltage v_o .
6. Repeat Problem 5 for the op-amp circuit shown in [Figure 6.41](#).

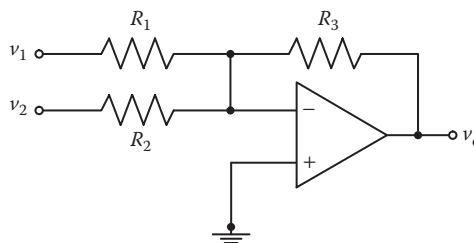


FIGURE 6.37
Problem 1.

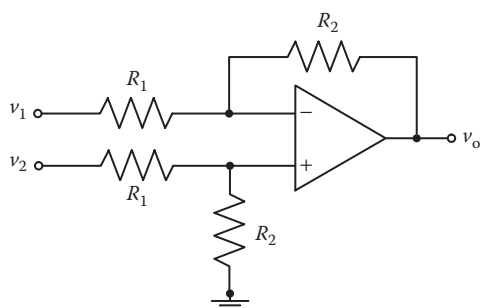


FIGURE 6.38
Problem 2.

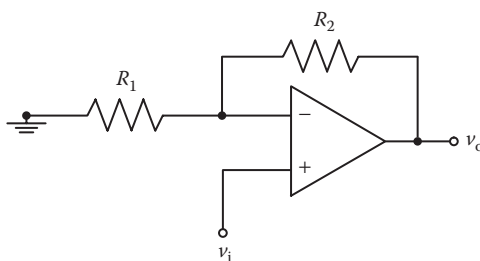


FIGURE 6.39
Problem 3.

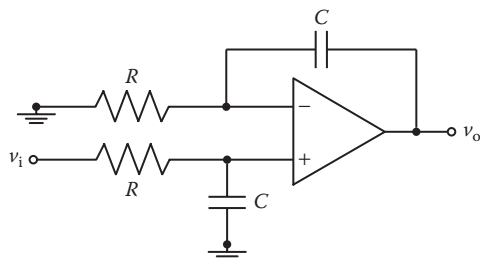


FIGURE 6.40
Problem 5.

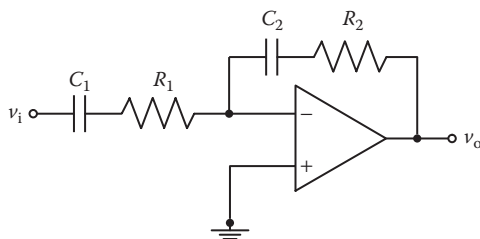


FIGURE 6.41
Problem 6.

6.4 Electromechanical Systems

Many useful devices, such as motors, generators, speakers, microphones, and accelerometers, are constructed by combining electrical elements and mechanical elements. For such electromechanical systems, we must apply electrical principles (e.g., Kirchhoff's laws) and mechanical principles (e.g., Newton's second law) to develop the dynamics model of the system. In this section, we discuss the modeling of direct current (DC) motors, which can generate forces or torques by using electrical subsystems and are essential actuators in control systems.

6.4.1 Elemental Relations of Electromechanical Systems

In a variety of electromechanical systems, electrical and mechanical subsystems are coupled by a magnetic field. Figure 6.42 shows a DC motor, which consists of basic elements (including the stator, the rotor, the armature, and the commutator). The stator provides a magnetic field across the rotor. The current is conducted to coils attached to the rotor via brushes, and the rotor is free to rotate. The combined unit of coils attached to the rotor is called the armature. The brushes are in contact with the rotating commutator, which causes the current to always be in the proper conductor windings, so as to produce a torque and keep it in the proper direction. The magnetic coupling relations between the electrical and mechanical subsystems in a DC motor can be derived using fundamental electromagnetic laws in introductory physics textbooks [4].

For simplicity, let us first consider a wire carrying a current within a magnetic field. Assume that the wire is either a straight conductor perpendicular to a uniform magnetic field or a circular conductor in a radial magnetic field. These are two common situations in many applications. Then, a force will be exerted on the wire, and the relation between the force f and the current i is

$$f = BLi \quad (6.25)$$

where:

B is the magnetic flux density in tesla ($1 \text{ T} = 1 \text{ Wb/m}^2$)

L is the length of the conductor in the magnetic field

The direction of the force can be determined using the right-hand rule, as shown in Figure 6.43a. Curl four fingers from the positive current direction to the positive direction of the magnetic field, and the thumb will point to the positive direction of the force.

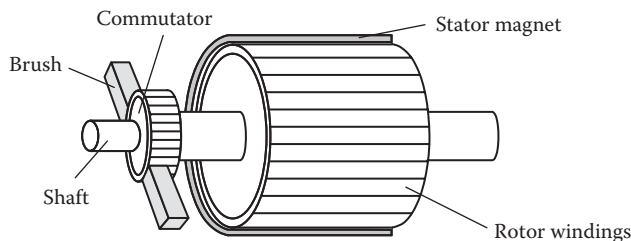
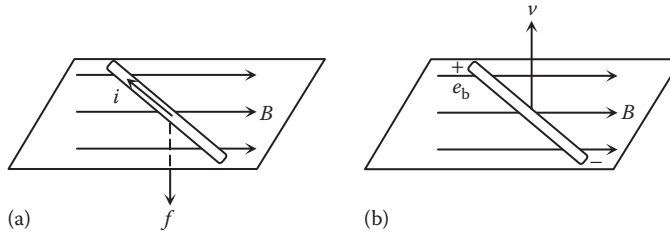


FIGURE 6.42
A DC brush motor.

**FIGURE 6.43**

The direction of (a) the force on a conductor and (b) the voltage induced in a moving conductor.

If the conductor moves relative to the magnetic field, a voltage will be induced in the conductor. Figure 6.43b shows a straight conductor moving upward in a magnetic field. Assume that the direction of the motion is perpendicular to the direction of the magnetic field. Then, the scalar relation between the induced voltage e_b and the velocity v of the conductor is

$$e_b = BLv \quad (6.26)$$

To avoid confusion, we use e to denote the voltage instead of v as we did in the previous sections. Again, using the right-hand rule, curl four fingers from the positive direction of the velocity to the positive field direction, and the thumb will point to the positive direction of the induced voltage e_b . Note that the induced voltage e_b opposes the current, and is known as back electromotive force (emf, an old term for voltage).

For the DC motor shown in Figure 6.42, assume that the armature current is i and the number of armature coils is n . By Equation 6.25, the force generated on the armature due to the magnetic field is $f = nBLi$. If the radius of the armature is r , the torque produced by the motor is

$$\tau_m = fr = nBLir = K_t i \quad (6.27)$$

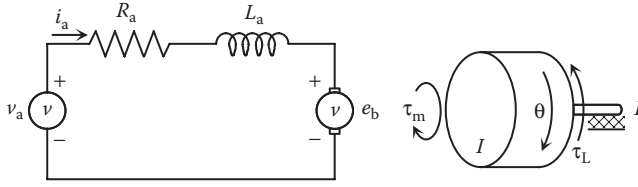
where $K_t = nBLr$ is the torque constant of the motor. Note that the linear velocity of the coils is proportional to the angular velocity, $v = \omega r$. Then, by Equation 6.26, the back emf generated in the armature due to the rotating motion is

$$e_b = nBLv = nBL\omega r = K_e \omega \quad (6.28)$$

where $K_e = nBLr$ is the back emf constant of the motor. The two constants, K_t and K_e , have the same expression and they will have the same numerical value if expressed in the same system of units. Equations 6.27 and 6.28 are used to model the coupling between the electrical and mechanical subsystems in a DC motor. Two primary types of DC motors, armature-controlled DC motors and field-controlled DC motors, are discussed next.

6.4.2 Armature-Controlled Motors

Figure 6.44 shows an electromechanical system with an armature-controlled DC motor. The electrical system is represented by an armature circuit, in which v_a is applied armature voltage, R_a is armature resistance, L_a is armature inductance, and e_b is back emf generated in the armature. The mechanical part is represented by a rotational system, in which I is

**FIGURE 6.44**

An electromechanical system with an armature-controlled DC motor.

the mass moment of inertia due to the rotor and the load, B is the viscous rotational damping associated with the load, τ_m is the torque produced by the motor, and τ_L is an additional torque applied to the load. In general, the load torque acts in the direction opposite to the motor torque. The differential equations of the system can be derived by using Kirchhoff's voltage law, the moment equation, and the electromechanical coupling relations.

For the electrical circuit, applying Kirchhoff's voltage law gives

$$\sum_j v_j = 0$$

$$R_a i_a + L_a \frac{di_a}{dt} + e_b - v_a = 0 \quad (6.29)$$

For the mechanical part, applying the moment equation gives

$$+ \curvearrowright: \sum M_C = I_C \alpha$$

$$\tau_m - \tau_L - B\dot{\theta} = I\ddot{\theta} \quad (6.30)$$

Substituting the coupling relations between the electrical and mechanical subsystems, $e_b = K_e \omega = K_e \dot{\theta}$ and $\tau_m = K_t i_a$, into Equations 6.29 and 6.30, we obtain

$$L_a \frac{di_a}{dt} + R_a i_a + K_e \dot{\theta} = v_a \quad (6.31)$$

$$I\ddot{\theta} + B\dot{\theta} - K_t i_a = -\tau_L \quad (6.32)$$

Note that the stiffness terms associated with the variable θ do not appear in Equations 6.31 and 6.32. Thus, the system dynamics can also be expressed in terms of ω instead of θ , as

$$L_a \frac{di_a}{dt} + R_a i_a + K_e \omega = v_a \quad (6.33)$$

$$I\dot{\omega} + B\omega - K_t i_a = -\tau_L \quad (6.34)$$

Assume that all the initial conditions are set to zero. Taking the Laplace transform of Equations 6.33 and 6.34 results in

$$L_a s I_a(s) + R_a I_a(s) = V_a(s) - K_e \Omega(s) \quad (6.35)$$

$$Is\Omega(s) + B\Omega(s) = -T_L(s) + K_t I_a(s) \quad (6.36)$$

Figure 6.45 shows a block diagram of the above-mentioned system. Using the basic rules for block diagram operation and reduction (see Section 4.5), we obtain the transfer function relating the armature voltage $v_a(t)$ and angular velocity $\omega(t)$, with $\tau_L = 0$,

$$\frac{\Omega(s)}{V_a(s)} = \frac{(1/(L_a s + R_a)) \cdot K_t \cdot (1/(Is + B))}{1 + (1/(L_a s + R_a)) \cdot K_t \cdot (1/(Is + B)) \cdot K_e} = \frac{K_t}{L_a Is^2 + (L_a B + R_a I)s + R_a B + K_t K_e} \quad (6.37)$$

The transfer function relating the load torque $\tau_L(t)$ and $\omega(t)$, with $v_a(t) = 0$, is

$$\frac{\Omega(s)}{T_L(s)} = \frac{-(1/(Is + B))}{1 - (1/(Is + B)) \cdot (-K_e) \cdot (1/(L_a s + R_a)) \cdot K_t} = -\frac{L_a s + R_a}{L_a Is^2 + (L_a B + R_a I)s + R_a B + K_t K_e} \quad (6.38)$$

Note that the two transfer functions have the same denominator, which is the characteristic of the system. The order of the characteristic polynomial implies that the system is second-order. The transfer functions $\Omega(s)/V_a(s)$ and $\Omega(s)/T_L(s)$ can also be obtained by solving Equations 6.35 and 6.36 by using Cramer's rule, and this is left as an exercise for the reader.

The motor model can be represented in state-space form by choosing appropriate states. We choose the armature current i_a and the angular velocity ω as the states. As shown in Figure 6.44, the armature voltage and the load torque are the two inputs. Let $u_1 = v_a$ and $u_2 = \tau_L$. Solving for the time derivatives di_a/dt and $\dot{\omega}$ from Equations 6.33 and 6.34 yields

$$\dot{x}_1 = \frac{di_a}{dt} = -\frac{R_a}{L_a} i_a - \frac{K_e}{L_a} \omega + \frac{1}{L_a} v_a \quad (6.39)$$

$$\dot{x}_2 = \dot{\omega} = \frac{K_t}{I} i_a - \frac{B}{I} \omega - \frac{1}{I} \tau_L \quad (6.40)$$

Thus, the state equation is

$$\begin{Bmatrix} \dot{x}_1 \\ \dot{x}_2 \end{Bmatrix} = \begin{bmatrix} -\frac{R_a}{L_a} & -\frac{K_e}{L_a} \\ \frac{K_t}{I} & -\frac{B}{I} \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix} + \begin{bmatrix} \frac{1}{L_a} & 0 \\ 0 & -\frac{1}{I} \end{bmatrix} \begin{Bmatrix} u_1 \\ u_2 \end{Bmatrix} \quad (6.41)$$

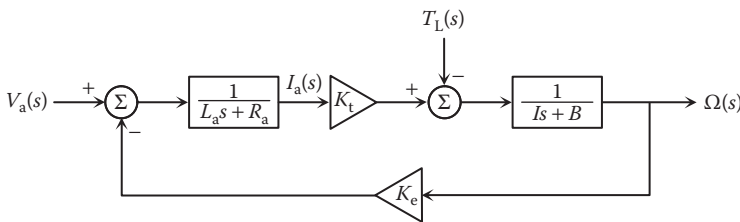


FIGURE 6.45

Block diagram of an armature-controlled DC motor.

If we select ω as the output, the output equation is

$$y = \begin{bmatrix} 0 & 1 \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix} + \begin{bmatrix} 0 & 0 \end{bmatrix} \begin{Bmatrix} u_1 \\ u_2 \end{Bmatrix} \quad (6.42)$$

Example 6.12: A Single-Link Robot Arm Driven by a Direct Current Motor

Consider the dynamic system shown in Figure 6.46, in which a single-link robot arm is driven by a DC motor. The differential equation of the robot arm in terms of the motor variable θ_m was determined in Example 5.20 to be

$$(I_m + N^2 I) \ddot{\theta}_m + (B_m + N^2 B) \dot{\theta}_m = \tau_m$$

where:

- I_m and I are the mass moments of inertia of the motor and the load, respectively
- B_m and B are the coefficients of the torsional viscous damping of the motor and the load, respectively
- τ_m is the torque generated by the motor
- N is the gear ratio

Assume that the armature inductance is negligibly small, that is, $L_a \approx 0$. The torque and the back emf constants of the motor are K_t and K_e , respectively.

- Derive the differential equation relating the applied voltage v_a and the link angular displacement θ .
- Determine the transfer function $\Theta(s)/V_a(s)$ by using the differential equation obtained in Part (a). Assume that all initial conditions are zero.

Solution

- For the electrical circuit, applying Kirchhoff's voltage law gives

$$R_a i_a + e_b - v_a = 0$$

where $e_b = K_e \dot{\theta}_m$. With the given gear ratio,

$$\theta = N \theta_m$$

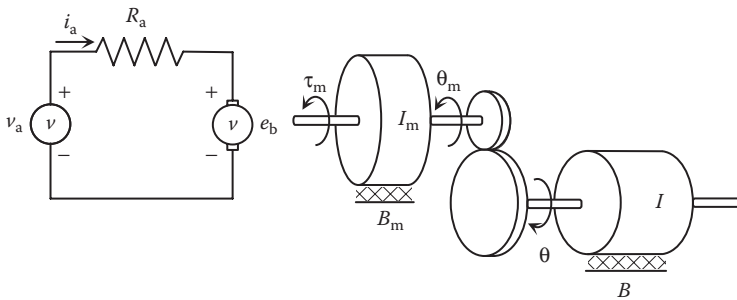


FIGURE 6.46

The model of a single-link robot arm driven by an armature-controlled DC motor.

we have

$$R_a i_a = v_a - K_e \frac{1}{N} \dot{\theta}$$

Thus, the current i_a can be expressed as

$$i_a = \frac{1}{R_a} v_a - \frac{K_e}{R_a} \frac{1}{N} \dot{\theta}$$

The model of the mechanical part in terms of θ is given by

$$(I_m + N^2 I) \frac{1}{N} \ddot{\theta} + (B_m + N^2 B) \frac{1}{N} \dot{\theta} = \tau_m$$

where $\tau_m = K_t i_a$. Substituting the expression of the current i_a , we obtain

$$(I_m + N^2 I) \frac{1}{N} \ddot{\theta} + (B_m + N^2 B) \frac{1}{N} \dot{\theta} = K_t \left(\frac{1}{R_a} v_a - \frac{K_e}{R_a} \frac{1}{N} \dot{\theta} \right)$$

Rearranging the equation yields

$$(I_m + N^2 I) \ddot{\theta} + \left(B_m + N^2 B + \frac{K_t K_e}{R_a} \right) \dot{\theta} = \frac{N K_t}{R_a} v_a$$

b. Taking the Laplace transform results in

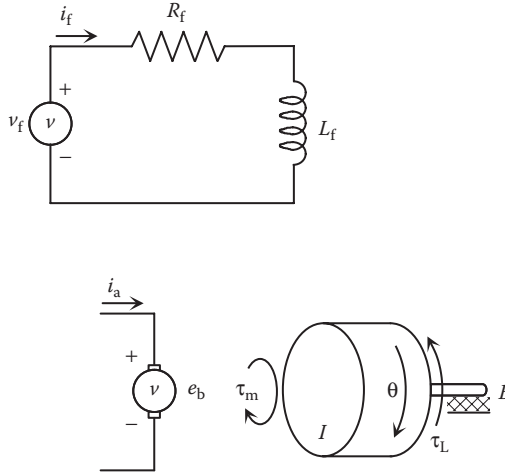
$$\frac{\Theta(s)}{V_a(s)} = \frac{N K_t / R_a}{(I_m + N^2 I) s^2 + (B_m + N^2 B + K_t K_e / R_a) s}$$

This electromechanical system can be modeled using Simulink and Simscape. The details will be discussed in [Section 6.6](#).

6.4.3 Field-Controlled Motors

In all but the smallest motors, the magnetic field is established by a current in separate field windings on the stator. For an armature-controlled DC motor, a constant current source is supplied to the field windings and the applied armature voltage, v_a , varies. Another way of controlling a DC motor is to keep the armature current i_a constant, while varying the voltage applied to the field windings. A simple model of a field-controlled DC motor is shown in [Figure 6.47](#), in which the shaft in the mechanical subsystem is assumed to be massless, rigid, and undamped. The electrical part is represented by a field circuit, where R_f is field resistance, L_f is field inductance, v_f is field voltage, and i_f is field current. Note that there is no back emf created in the field circuit. The torque generated by the motor is proportional to the field current,

$$\tau_m = K_t i_f \quad (6.43)$$

**FIGURE 6.47**

An electromechanical system with a field-controlled DC motor.

The system under consideration has two inputs, v_f and τ_L . Two independent variables, i_f and θ , can be used to describe the system dynamics. For the electrical part, we apply Kirchhoff's voltage law to the field circuit,

$$L_f \frac{di_f}{dt} + R_f i_f = v_f \quad (6.44)$$

For the mechanical part, introducing the motor equation (Equation 6.43) and applying the moment equation gives

$$I\ddot{\theta} + B\dot{\theta} - K_t i_f = -\tau_L \quad (6.45)$$

or

$$I\dot{\omega} + B\omega - K_t i_f = -\tau_L \quad (6.46)$$

Equations 6.44 and 6.45, or Equations 6.44 and 6.46, are the system differential equations of the field-controlled DC motor.

Assuming zero initial conditions and taking the Laplace transform of Equations 6.44 and 6.46, we have

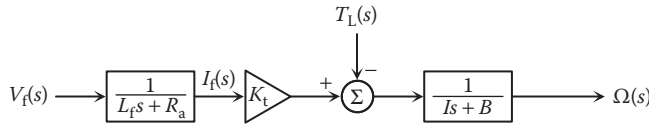
$$L_f s I_f(s) + R_f I_f(s) = V_f(s) \quad (6.47)$$

$$Is\Omega(s) + B\Omega(s) = K_t I_f(s) - T_L(s) \quad (6.48)$$

A block diagram of the system is shown in [Figure 6.48](#).

The transfer functions $\Omega(s)/V_f(s)$ and $\Omega(s)/T_L(s)$ can be easily derived from the diagram,

$$\frac{\Omega(s)}{V_f(s)} = \frac{K_t}{(L_f s + R_f)(Is + B)} = \frac{K_t}{L_f Is^2 + (L_f B + R_f I)s + R_f B} \quad (6.49)$$

**FIGURE 6.48**

Block diagram of a field-controlled DC motor.

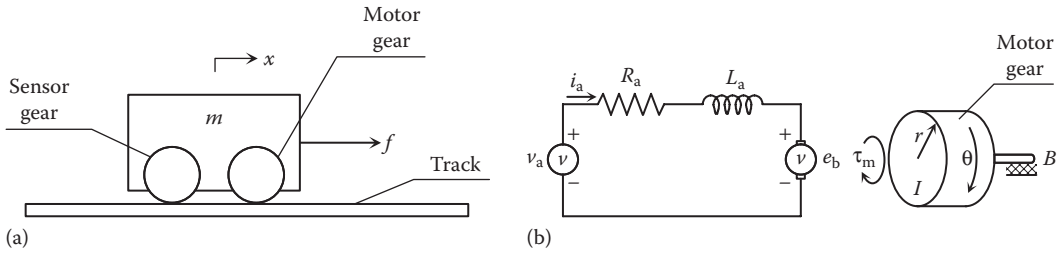
$$\frac{\Omega(s)}{T_L(s)} = -\frac{1}{Is + B} \quad (6.50)$$

If we choose the field current and the angular velocity as the state variables ($x_1 = i_f$ and $x_2 = \omega$), the field voltage and the load torque as the inputs ($u_1 = v_f$ and $u_2 = \tau_L$), and the angular velocity as the output ($y = \omega$), the state-space form is

$$\begin{Bmatrix} \dot{x}_1 \\ \dot{x}_2 \end{Bmatrix} = \begin{bmatrix} -\frac{R_f}{L_f} & 0 \\ \frac{K_t}{I} & -\frac{B}{I} \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix} + \begin{bmatrix} \frac{1}{L_f} & 0 \\ 0 & -\frac{1}{I} \end{bmatrix} \begin{Bmatrix} u_1 \\ u_2 \end{Bmatrix}, \quad y = \begin{bmatrix} 0 & 1 \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix} + \begin{bmatrix} 0 & 0 \end{bmatrix} \begin{Bmatrix} u_1 \\ u_2 \end{Bmatrix} \quad (6.51)$$

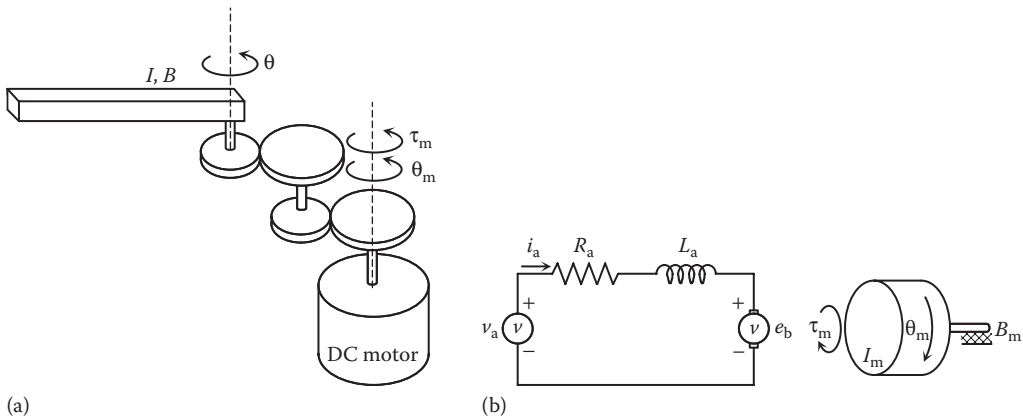
PROBLEM SET 6.4

- Reconsider the armature-controlled motor in [Figure 6.44](#). Equations 6.31 and 6.32 represent the dynamics of the system in terms of the variables i_a and θ .
 - Assuming the angle θ to be the output, draw a block diagram to represent the dynamics of the armature-controlled motor.
 - Derive the transfer functions $\Theta(s)/V_a(s)$ and $\Theta(s)/T_L(s)$. All of the initial conditions are assumed to be zero.
 - Determine the state-space form.
- Reconsider the field-controlled motor in [Figure 6.47](#). Equations 6.44 and 6.45 represent the dynamics of the system in terms of the variables i_f and θ .
 - Assuming the angle θ to be the output, draw a block diagram to represent the dynamics of the field-controlled motor.
 - Derive the transfer functions $\Theta(s)/V_f(s)$ and $\Theta(s)/T_L(s)$. All of the initial conditions are assumed to be zero.
 - Determine the state-space form.
- Consider the electromechanical system shown in [Figure 6.49a](#). It consists of a cart of mass m moving without slipping on a ground track. The cart is equipped with an armature-controlled DC motor, which is coupled to a rack and pinion mechanism to convert the rotational motion to translation and to create the driving force f for the system. [Figure 6.49b](#) shows the equivalent electric circuit and the mechanical model of the DC motor, where r is the radius of the motor gear. The torque and the back emf constants of the motor are K_t and K_e , respectively.
 - Derive the differential equation of the system relating the cart position x and the applied voltage v_a .
 - Determine the transfer function $X(s)/V_a(s)$ by using the differential equation obtained in Part (a). Assume that all initial conditions are zero.

**FIGURE 6.49**

Problem 3. (a) A DC motor-driven cart and (b) the equivalent electric circuit and the mechanical model of the DC motor.

4. Consider the single-link robot arm, as shown in Figure 6.50a. It is driven by an armature-controlled DC motor through spur gears with a total gear ratio of N and $\theta = N\theta_m$. The mass moments of inertia of the motor and the load are I_m and I , respectively. The coefficients of torsional viscous damping of the motor and the load are B_m and B , respectively. Figure 6.50b shows the equivalent electric circuit and the mechanical model of the DC motor. The torque and the back emf constants of the motor are K_t and K_e , respectively.
 - a. Determine the transfer function $\Theta(s)/V_a(s)$. Assume that all initial conditions are zero.
 - b. Determine the differential equation relating the applied voltage v_a and the link angular displacement θ .
5. A more complicated model of the armature-controlled motor is shown in Figure 6.51, in which the rotor is connected to an inertial load through a flexible and damped shaft. K_m and B_m represent the torsional stiffness and the torsional viscous damping of the shaft, respectively. The mass moments of inertia of the motor and the load are I_m and I_L , respectively. Let $\omega_m = \dot{\theta}_m$ and $\omega_L = \dot{\theta}_L$.
 - a. Assuming zero initial conditions, derive the transfer functions $\Omega_L(s)/V_a(s)$ and $\Omega_L(s)/T_L(s)$.

**FIGURE 6.50**

Problem 4. (a) A DC motor-driven single-link robot arm and (b) the equivalent electric circuit and the mechanical model of the DC motor.

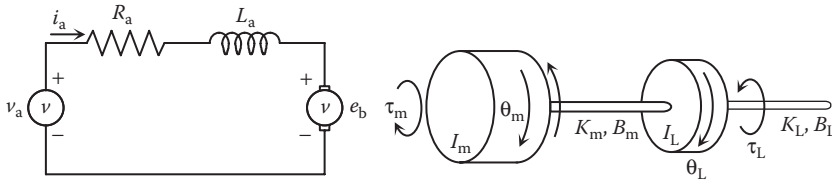


FIGURE 6.51
Problem 5.

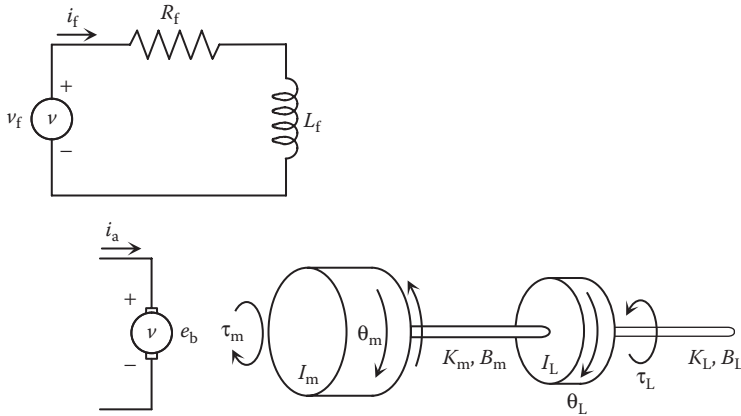


FIGURE 6.52
Problem 6.

- b. Assuming the angular velocity ω_L to be the output, draw a block diagram to represent the dynamics of the armature-controlled motor.
 - c. Determine the state-space form.
6. A more complicated model of the field-controlled motor is shown in Figure 6.52, in which the rotor is connected to an inertial load through a flexible and damped shaft. K_m and B_m represent the torsional stiffness and the torsional viscous damping of the shaft, respectively. The mass moments of inertia of the motor and the load are I_m and I_L , respectively.
- a. Assuming zero initial conditions, derive the transfer functions $\Omega_L(s)/V_f(s)$ and $\Omega_L(s)/T_L(s)$.
 - b. Assuming the angular velocity ω_L to be the output, draw a block diagram to represent the dynamics of the field-controlled motor.
 - c. Determine the state-space form.

6.5 Impedance Methods

The concept of impedance is very useful in electrical systems because it provides an alternative to transfer functions and differential equations for the derivation of system mathematical models.

6.5.1 Impedances of Electric Elements

Impedance is a generalization of the concept of resistance. Mathematically, electrical impedance is defined as the ratio of the voltage to the current in the s domain,

$$Z(s) = \frac{V(s)}{I(s)} \quad (6.52)$$

For a resistor, we have $v = Ri$, hence the impedance is its resistance R ,

$$Z(s) = R \quad (6.53)$$

For an inductor, we have $v = Ldi/dt$. Assuming zero initial conditions, this yields $V(s) = LsI(s)$. Thus, the impedance of an inductor is

$$Z(s) = Ls \quad (6.54)$$

Similarly, for a capacitor, $i = Cdv/dt$ yields $I(s) = CsV(s)$. Thus, the impedance of a capacitor is

$$Z(s) = \frac{1}{Cs} \quad (6.55)$$

6.5.2 Series and Parallel Impedances

Because impedance can be viewed as a generalized resistance, it is easy to find the equivalent impedance for series-connected or parallel-connected electrical elements. [Figure 6.53](#) shows n impedances in series. Note that the same current flows through n impedances and the total voltage drop across them is

$$V(s) = I(s)Z_1(s) + I(s)Z_2(s) + \cdots + I(s)Z_n(s) \quad (6.56)$$

In the equivalent diagram, the relation between the current and the voltage is

$$V(s) = I(s)Z_{\text{eq}}(s) \quad (6.57)$$

Thus,

$$Z_{\text{eq}}(s) = Z_1(s) + Z_2(s) + \cdots + Z_n(s) \quad (6.58)$$

That is, the equivalent impedance Z_{eq} is equal to the sum of all the individual impedances Z_i .

If there are n impedances in parallel, as shown in [Figure 6.54](#), all the impedances have the same voltage drop. The total current through all the elements is

$$I(s) = \frac{V(s)}{Z_1(s)} + \frac{V(s)}{Z_2(s)} + \cdots + \frac{V(s)}{Z_n(s)} \quad (6.59)$$

For the equivalent impedance,

$$I(s) = \frac{V(s)}{Z_{\text{eq}}(s)} \quad (6.60)$$

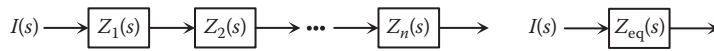


FIGURE 6.53
Equivalence for impedance in series.

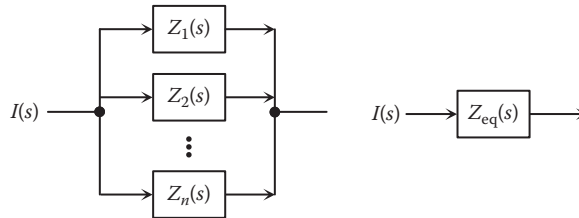


FIGURE 6.54
Equivalence for impedances in parallel.

Therefore,

$$\frac{1}{Z_{\text{eq}}(s)} = \frac{1}{Z_1(s)} + \frac{1}{Z_2(s)} + \cdots + \frac{1}{Z_n(s)} \quad (6.61)$$

That is, the reciprocal of the equivalent impedance Z_{eq} is equal to the sum of all the reciprocals of the individual impedances Z_i .

Note that impedance is essentially a transfer function, which has no integral or derivative signs. If we redraw an electrical system in the s domain by replacing passive elements with their corresponding impedances, we can determine the transfer function of the system by using Kirchhoff's laws along with series and parallel laws. The differential equation of the system can then be obtained by converting the transfer function back from the s domain to the time-domain. Thus, the concept of impedance provides another way of modeling electrical systems, without writing any time-domain equations.

Example 6.13: Impedance Method: An RLC Circuit

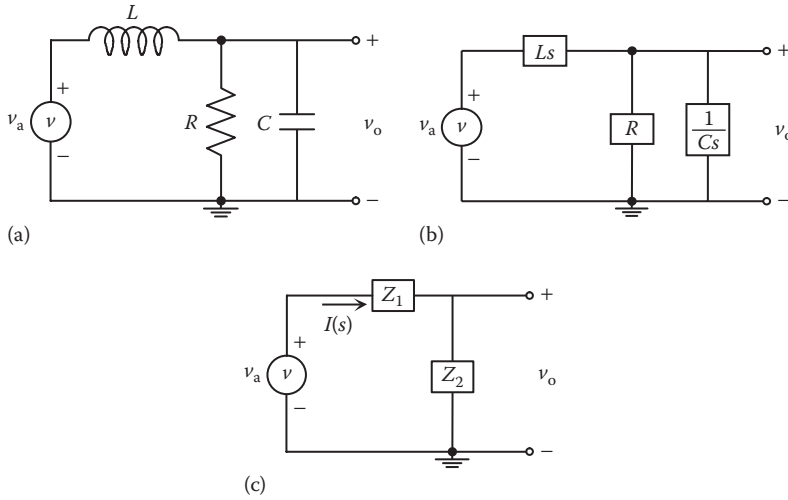
For the electric circuit in Example 6.4, use the impedance method to derive the differential equation relating the output voltage $v_o(t)$ to the input voltage $v_a(t)$. Assume zero initial conditions.

Solution

The original electric circuit is shown in Figure 6.55a. We can replace the passive elements with their impedance representations and redraw the circuit in the s domain, as shown in Figure 6.55b. Note that the resistor R is in parallel connection with the capacitor C . The corresponding equivalent impedances are

$$Z_1(s) = Ls$$

$$\frac{1}{Z_2(s)} = \frac{1}{R} + \frac{1}{1/(Cs)}$$

**FIGURE 6.55**

An RLC circuit drawn (a) in time domain, (b) in s domain, and (c) using impedances.

or

$$Z_2(s) = \frac{R}{RCs + 1}$$

For the equivalent impedance circuit in [Figure 6.55c](#), we apply Kirchhoff's voltage law,

$$Z_1(s)I(s) + Z_2(s)I(s) - V_a(s) = 0$$

where the current is related to the output voltage by

$$V_o(s) = Z_2(s)I(s)$$

Thus, we have

$$Z_1(s) \frac{V_o(s)}{Z_2(s)} + V_o(s) = V_a(s)$$

which gives the transfer function relating the input voltage v_a and the output voltage v_o ,

$$\frac{V_o(s)}{V_a(s)} = \frac{Z_2(s)}{Z_1(s) + Z_2(s)} = \frac{R}{RLCs^2 + Ls + R}$$

or

$$(RLCs^2 + Ls + R)V_o(s) = RV_a(s)$$

By taking the inverse Laplace transform with the assumption of zero initial conditions, we obtain the differential equation of the system

$$RLC\ddot{v}_o + L\dot{v}_o + Rv_o = Rv_a$$

which is the same as the one obtained in Example 6.4.

Example 6.14: Impedance Method: An Op-Amp Circuit

For the op-amp circuit in Example 6.11, use the impedance method to derive the differential equation relating the output voltage $v_o(t)$ to the input voltage $v_i(t)$. Assume zero initial conditions.

Solution

The original op-amp circuit is shown in Figure 6.56a. Replacing the passive elements with their impedance representations gives the equivalent op-amp circuit, as shown in Figure 6.56b, where

$$\frac{1}{Z_1(s)} = \frac{1}{R_1} + \frac{1}{1/(C_1s)}$$

$$\frac{1}{Z_2(s)} = \frac{1}{R_2} + \frac{1}{1/(C_2s)}$$

or

$$Z_1(s) = \frac{R_1}{R_1C_1s + 1}$$

$$Z_2(s) = \frac{R_2}{R_2C_2s + 1}$$

Because the current drawn by the op-amp is very small, applying Kirchhoff's current law to node 1 yields,

$$I_1(s) = I_2(s)$$

$$\frac{V_i(s) - V_1(s)}{Z_1(s)} = \frac{V_1(s) - V_o(s)}{Z_2(s)}$$

where the voltage at node 1 obeys

$$V_1(s) = V_-(s) = V_+(s) = 0$$

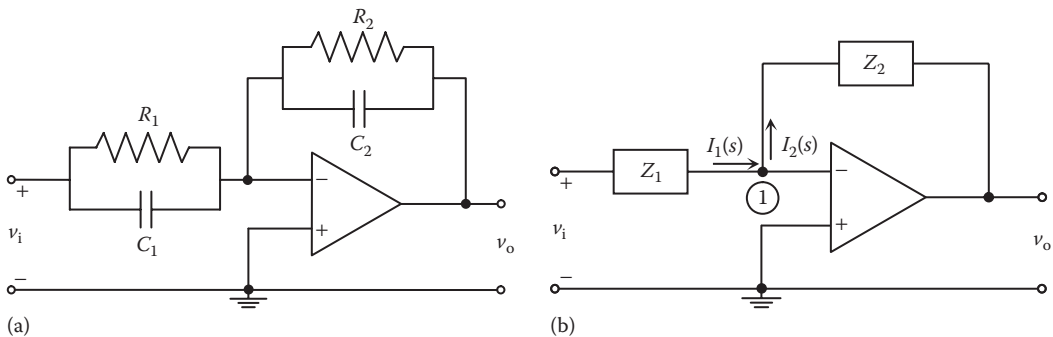


FIGURE 6.56

An op-amp circuit drawn (a) in time domain and (b) using impedances.

Consequently, we have

$$\frac{V_i(s)}{Z_1(s)} = -\frac{V_o(s)}{Z_2(s)}$$

which gives the transfer function relating the input voltage v_i and the output voltage v_o ,

$$\frac{V_o(s)}{V_i(s)} = -\frac{Z_2(s)}{Z_1(s)} = -\frac{R_1 R_2 C_1 s + R_2}{R_1 R_2 C_2 s + R_1}$$

or

$$(R_1 R_2 C_2 s + R_1) V_o(s) = -(R_1 R_2 C_1 s + R_2) V_i(s)$$

By taking the inverse transform with the assumption of zero initial conditions, we obtain the differential equation of the system

$$R_1 R_2 C_2 \dot{v}_o + R_1 v_o = -R_1 R_2 C_1 \dot{v}_i - R_2 v_i$$

which is the same as the one obtained in Example 6.11.

6.5.3 Mechanical Impedances

Analogous to electrical impedance, mechanical impedance is defined as

$$Z(s) = \frac{V(s)}{F(s)} \quad (6.62)$$

where $V(s)$ and $F(s)$ are the Laplace transforms of velocity $v(t)$ and force $f(t)$, respectively. The impedance concept can also be used to obtain models of mechanical systems along with the linear graph, which is a topic beyond the scope of this text. Here, we only give the definitions of impedances for fundamental mechanical elements.

For a viscous damper, the damping force is related to the velocity by $f = bv$ or $F(s) = bV(s)$. Thus, the impedance of a damper is

$$Z(s) = \frac{1}{b} \quad (6.63)$$

For a spring element, the spring force is proportional to the displacement, $f = kx = k \int v dt$ or $F(s) = kV(s)/s$. Thus,

$$Z(s) = \frac{s}{k} \quad (6.64)$$

For a mass element, by Newton's second law, $f = ma = m\dot{v}$ or $F(s) = msV(s)$. Thus, the impedance of a mass element is

$$Z(s) = \frac{1}{ms} \quad (6.65)$$

When comparing the two sets of equations, Equations 6.53 through 6.55 and 6.63 through 6.65, we note that the corresponding electrical and mechanical elements are not equivalent, although they have similar physical effects. For example, both the resistor and the damper dissipate energy. But the mathematical expressions for their impedances are different.

PROBLEM SET 6.5

1. Reconsider the RC circuit shown in [Figure 6.24](#). Use the impedance method to determine the transfer function $I(s)/V_a(s)$ and the input–output differential equation relating v_c and v_a . Assume that all the initial conditions are zero.
2. Reconsider the RL circuit shown in [Figure 6.25](#). Use the impedance method to determine the transfer function $V_L(s)/V_a(s)$ and the input–output differential equation relating i_L and v_a . Assume zero initial conditions.
3. Reconsider the RLC circuit shown in [Figure 6.26](#). Use the impedance method to determine the input–output differential equation relating v_o and v_a . Assume zero initial conditions.
4. Reconsider the RLC circuit shown in [Figure 6.27](#). Use the impedance method to determine the input–output differential equation relating i and v_a . Assume that all the initial conditions are zero.
5. Reconsider the RLC circuit shown in [Figure 6.28](#). Use the impedance method to determine the input–output differential equation relating v_o and v_a . Assume zero initial conditions.
6. Reconsider the RLC circuit shown in [Figure 6.29](#). Use the impedance method to determine the input–output differential equation relating v_o and v_a . Assume zero initial conditions.
7. Reconsider the RL circuit shown in [Figure 6.30](#). Use the impedance method to determine the input–output differential equation relating v_o and v_a . Assume that all initial conditions are zero.
8. Reconsider the op-amp circuit shown in [Figure 6.41](#). Use the impedance method to determine the differential equation relating the input voltage v_i and the output voltage v_o .

6.6 System Modeling with Simulink and Simscape

Electrical systems, or electrical circuits, can be modeled as systems with interconnected electrical elements, such as resistors, inductors, capacitors, op-amps, and others. These passive electrical elements are connected with active electrical elements, including current sources and voltage sources. The common output signals of electrical systems are currents and voltages. Similar to modeling of mechanical systems, the dynamics of electrical systems can be represented by ordinary differential equations, transfer functions, or the state-space form. Therefore, the Simulink modeling techniques discussed in [Section 5.6](#) can also be applied to electrical systems.

This section focuses on physical modeling of electrical systems with Simscape. The blocks in the library of Simscape\Foundation Library\Electrical are categorized into three types, Electrical Elements, Electrical Sources, and Electrical Sensors. The Simscape library of Electrical Elements includes basic electrical building blocks, such as resistors, inductors, capacitors, op-amps, electromechanical convertors, switches, grounds, etc. The Simscape library of Electrical sources includes DC, AC, and controlled voltage and current sources. The Simscape library of Electrical Sensors includes two types of sensors, current and voltage sensors. The examples in this section illustrate Simscape modeling of electric circuits, op-amp circuits, and DC motors.

6.6.1 Electric Circuits

To correctly model an electric circuit by using Simscape, it is very important to verify polarity and connection to ground. First, the + and – signs seen on the ports of a block indicate how the current flows through that block. Second, each topologically distinct electric circuit must contain at least one Electrical Reference block, which represents a connection to the ground.

Example 6.15: A Series RLC Circuit

Consider the series RLC circuit shown in [Figure 6.57](#), where $R = 1\ \Omega$, $L = 1\ \text{H}$, and $C = 0.5\ \text{F}$. When the switch is closed at 1 second, the circuit is driven by a 24 V DC voltage source. Assume that all initial conditions are zero.

- Build a Simscape model of the physical system and find the loop current $i(t)$ and the voltage across the capacitor $v_C(t)$.
- Refer to the results obtained in Example 6.1. Build a Simulink model of the system based on the transfer function $I(s)/V_a(s)$ and find the loop current $i(t)$.
- Refer to the results obtained in Example 6.1. Build a Simulink model of the system based on the transfer function $V_C(s)/V_a(s)$ and find the voltage across the capacitor $v_C(t)$.

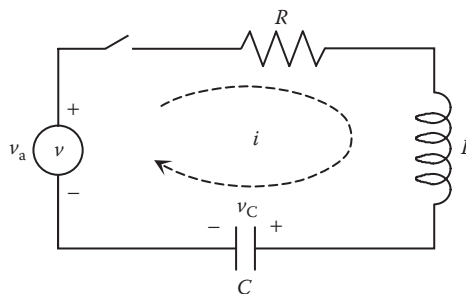
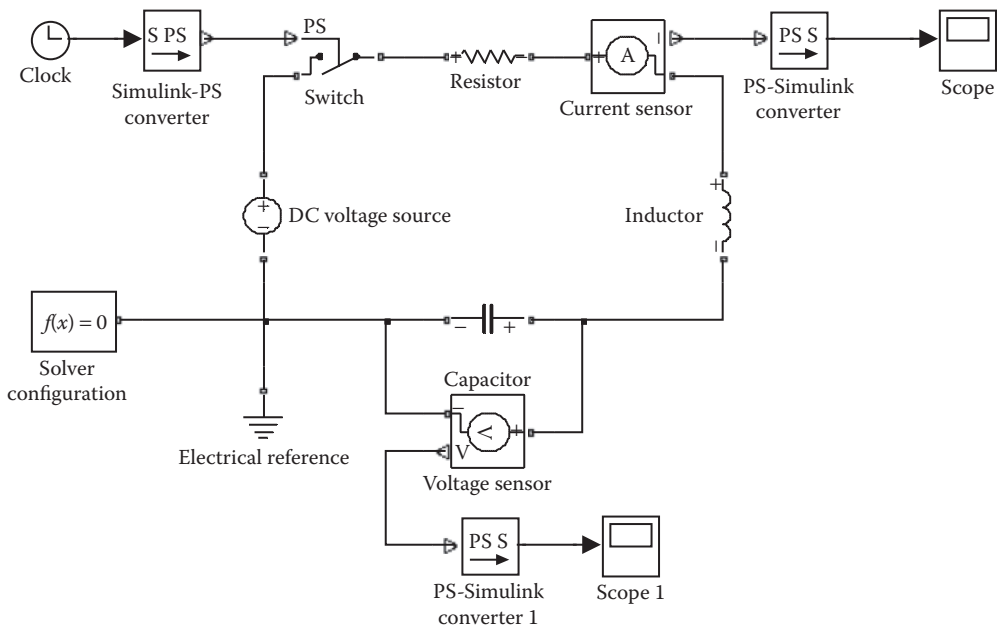


FIGURE 6.57

A series RLC circuit with a switch closed at 1 second.

Solution

- a. The Simscape block diagram corresponding to the physical system is shown in [Figure 6.58](#), which can be created by following these steps.
1. Type `ssc_new` at the MATLAB Command window to open the main Simscape library and create a new model.
 2. Open the library of Simscape/Foundation Library/Electrical/Electrical Elements and drag the Resistor, Inductor, and Capacitor blocks into the model window. Double-click these blocks to define the parameters Resistance, Inductance, and Capacitance as 1 Ohm, 1 H, and 0.5 F. Also, drag the Switch and Electrical Reference blocks into the model window.
 3. To add a 24 V DC voltage source, open the library of Simscape/Foundation Library/Electrical/Electrical Sources and drag the DC Voltage Source block into the model window. Double-click the block and define the parameter Constant voltage as 24 V.
 4. To measure the loop current and the voltage across the capacitor, open the library of Simscape/Foundation Library/Electrical/Electrical Sensors and drag both the Current Sensor and Voltage Sensor blocks into the model window. Each sensor has ports + and –, through which the sensor is connected to the circuit. The third port, I for the Current Sensor or V for the Voltage Sensor, is a physical signal port that outputs either current or voltage value.

**FIGURE 6.58**

Simscape block diagram corresponding to Example 6.15.

5. Note that the Switch is closed at 1 second. To add the time source, open the library of Simulink/Sources and drag the Clock block, which is connected to the Simulink-PS Converter block and then the Switch block. Double-click the Switch block and type 1 for the Threshold. This implies that the switch is closed if the time is greater than 1 second, otherwise the switch is open.
 6. To display the loop current and the voltage across the capacitor, open the library of Simulink/Sinks and drag two Scope blocks, which are connected to the sensor blocks through the PS-Simulink Converter blocks. Note that the Current Sensor is connected in series with the circuit and the Voltage Sensor is connected in parallel with the Capacitor block.
 7. Orient the blocks and connect them as shown in Figure 6.58. Set the simulation time to 15 seconds and run the model. The plots of the resulting loop current $i(t)$ and the voltage across the capacitor $v_c(t)$ are shown in Figures 6.59 and 6.60, respectively.
- b. Refer to the results obtained in Example 6.1. The transfer function relating the input $v_a(t)$ to the output $i(t)$ is

$$\frac{I(s)}{V_a(s)} = \frac{Cs}{LCs^2 + RCs + 1}$$

When the switch is closed at 1 second, the circuit is driven by a 24V DC voltage source. Mathematically, this can be modeled using a Step block with the Step time set as 1 and the Final value set as 24. The corresponding Simulink block diagram is shown in Figure 6.61, where a Transfer Fcn block is used to represent the series RLC circuit. Double-click the block and type [C 0] for the Numerator coefficient and [L*C R*C 1] for the Denominator coefficient to define the transfer function $I(s)/V_a(s)$.

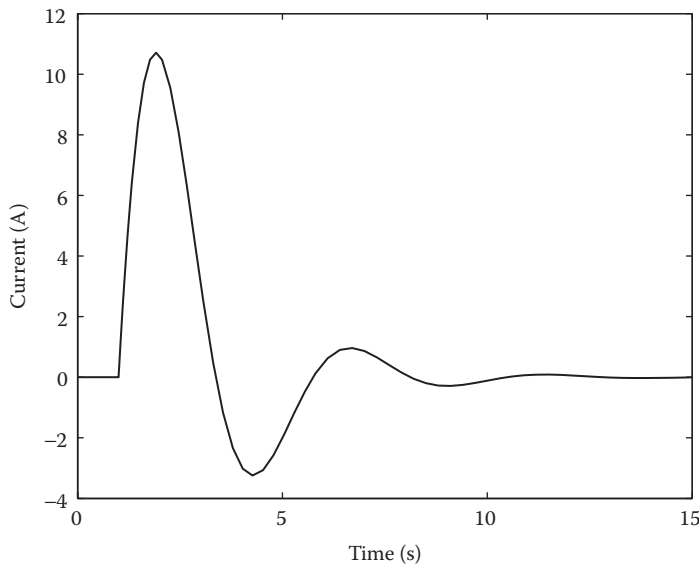


FIGURE 6.59

Loop current output $i(t)$ of the series RLC circuit in Example 6.15.

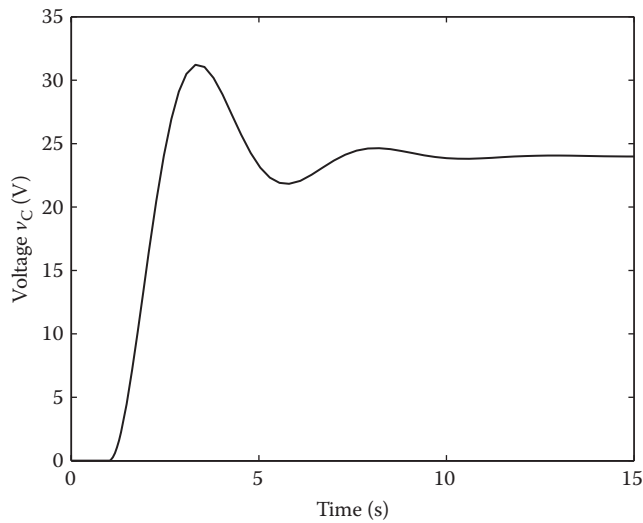


FIGURE 6.60
Voltage across the capacitor $v_C(t)$ in Example 6.15.

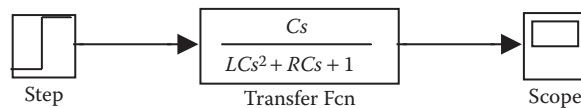


FIGURE 6.61
Simulink block diagram built based on the transfer function $I(s)/V_a(s)$.

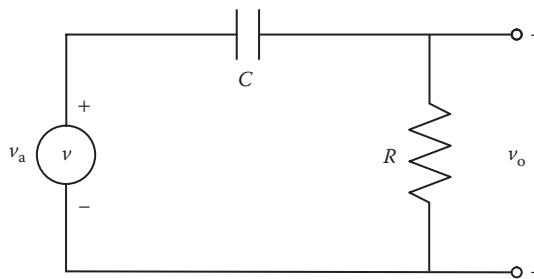
- c. Similarly, the transfer function relating the input $v_a(t)$ to the output $v_C(t)$ is obtained as

$$\frac{V_C(s)}{V_a(s)} = \frac{1}{LCs^2 + RCs + 1}$$

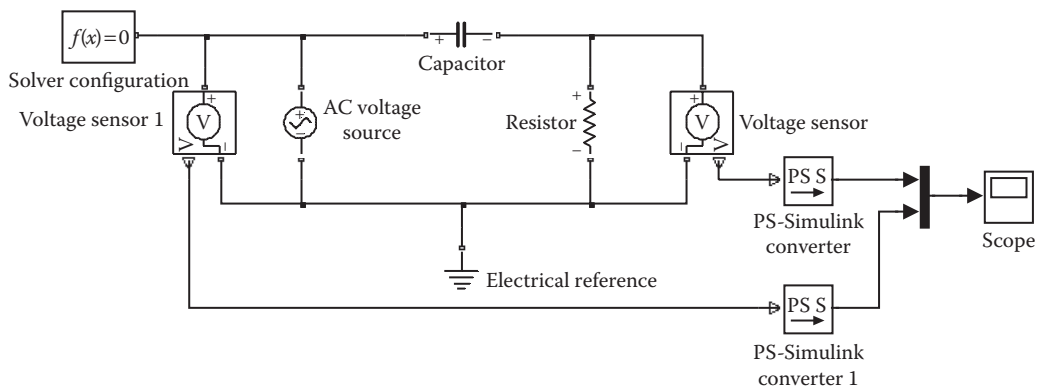
The corresponding Simulink block diagram is similar to the one shown in [Figure 6.61](#), except that the Numerator coefficient of the Transfer Fcn block is $[1]$, instead of $[C \ 0]$. Running the Simulink models, we will obtain the same current output as shown in [Figure 6.59](#) and the voltage output as shown in [Figure 6.60](#).

Example 6.16: An RC High-Pass Filter

A passive, analog, first-order high-pass filter can be realized by an RC circuit (see [Figure 6.62](#)), which passes high-frequency signals but attenuates signals at low frequencies. Assume that the resistance is $R = 100 \ \Omega$ and the capacitance is $C = 10 \ \mu\text{F}$. The circuit is connected to an AC voltage source, which has an amplitude of 1 V and a frequency varying from 1 to 1000 Hz. Build a Simscape model of the physical system and find the output voltage $v_o(t)$ when the frequency of the input voltage is 1, 10, 100, and 1000 Hz.

**FIGURE 6.62**

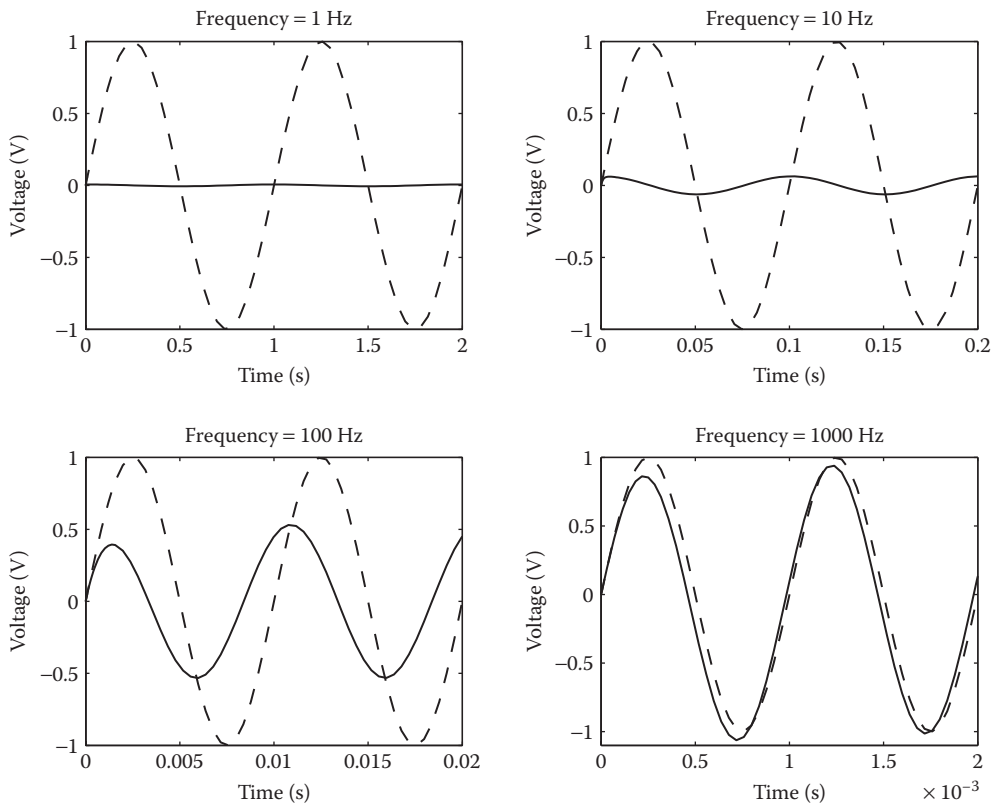
A passive, analog, first-order high-pass filter realized by an RC circuit.

**FIGURE 6.63**

Simscape block diagram corresponding to Example 6.16.

Solution

The Simscape block diagram of the RC high-pass filter is shown in [Figure 6.63](#), where the AC Voltage Source block outputs a sinusoidal voltage. To better understand high-pass filtering, two Voltage Sensor blocks are included to measure the input and output voltages. They are displayed on the same scope through a Mux block, which can be found in the Simulink library of Signal Routing. Double-click the AC Voltage Source block, type 1 for the Peak amplitude, and choose the unit as V. Vary the Frequency from 1, 10, 100, to 1000 Hz and run the simulations. The comparison between the input voltage and output voltage is given in [Figure 6.64](#), in which solid lines are the output voltages and the dashed ones are the input voltages. It is obvious that the filter passes high-frequency signals but attenuates signals at low frequencies. The reader can derive the transfer function of the system and build a Simulink block diagram as an exercise.

**FIGURE 6.64**

Comparison between the input and output voltages of the high-pass filter in Example 6.16.

6.6.2 Operational Amplifiers

The Op-Amp block in the Simscape library of Electrical Elements models the ideal behavior of an op-amp. The block has three electrical conserving ports. As discussed in [Section 6.3](#), the voltage at the positive pin is equal to the voltage at the negative pin. In other words, the op-amp gain is assumed to be infinite. This implies that the current from the positive terminal to the negative terminal is zero. Building an op-amp circuit by using the Op-Amp block, combined with other electrical elements, such as resistors, capacitors, and so on, is very straightforward.

Example 6.17: An Op-Amp Differentiator

Consider the op-amp differentiator in Example 6.10. Assume that the parameter values are $R = 1 \text{ M}\Omega$ and $C = 1 \text{ }\mu\text{F}$. Build a Simscape model of the op-amp circuit and find the output voltage $v_o(t)$ when the input voltage is $v_i = -0.1t \text{ V}$.

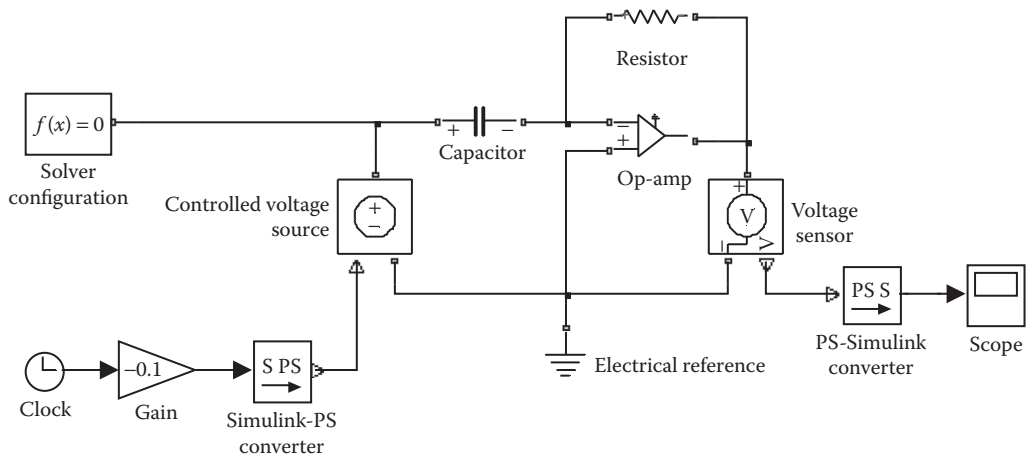


FIGURE 6.65
Simscape block diagram corresponding to Example 6.17.

Solution

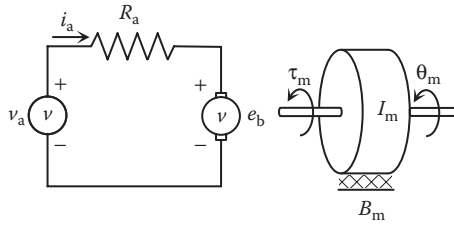
The Simscape block diagram of the op-amp differentiator is shown in Figure 6.65, in which a Clock block and a Gain block are used to generate the input voltage $v_i = -0.1t$ V. The Simulink input signal is converted into an equivalent voltage source by using a Controlled Voltage Source block. According to the result in Example 6.10, the output voltage v_o is proportional to the time derivative of the input voltage v_i via $v_o = -RC\dot{v}_i$. Substitute the parameter values, and a constant value of 0.1 is expected to be the output voltage. The problem of building a Simulink block diagram based on the differential equation is left to the reader as an exercise.

6.6.3 Direct Current Motors

As presented in Section 6.4, the mathematical model of a DC motor can be represented using a set of differential equations, a set of transfer functions, or in the state-space form. For each representation, a Simulink model can be built using the corresponding blocks, such as Transfer Fcn and State-Space blocks. To build a Simscape model of a DC motor, which is an electromechanical system, we need to use the blocks in both the Simscape\Foundation Library\Electrical and the Simscape\Foundation Library\Mechanical libraries. Two blocks, Translational Electromechanical Converter and Rotational Electromechanical Converter, in the Simscape library of Electrical Elements provide an interface between the electrical and mechanical translational or rotational domains. The following example illustrates how to model an armature-controlled DC motor with Simulink and Simscape.

Example 6.18: An Armature-Controlled Direct Current Motor

Consider the dynamic system shown in Figure 6.66, which represents an armature-controlled DC motor. Assume that the armature inductance is negligibly small, that is, $L_a = 0$. The system dynamics can be expressed as

**FIGURE 6.66**

An armature-controlled DC motor with negligible inductance.

$$R_a i_a + K_e \dot{\theta}_m = v_a$$

$$I_m \ddot{\theta}_m + B_m \dot{\theta}_m - K_t i_a = 0$$

where the armature resistance is $R_a = 0.5 \, \Omega$, the back emf constant is $K_e = 0.05 \, \text{V}\cdot\text{s}/\text{rad}$, the torque constant is $K_t = 0.05 \, \text{N}\cdot\text{m}/\text{A}$, the mass moment of inertia of the motor is $I_m = 0.00025 \, \text{kg}\cdot\text{m}^2$, the coefficient of the torsional viscous damping of the motor is $B_m = 0.0001 \, \text{N}\cdot\text{m}\cdot\text{s}/\text{rad}$, and the applied voltage is $v_a = 10 \, \text{V}$.

- Denote $\omega_m = \dot{\theta}_m$. Following [Figure 6.45](#), build a Simulink block diagram by using the given equations and find the armature current output $i_a(t)$ and the rotor speed $\omega_m(t)$.
- Assume zero initial conditions, determine the transfer functions $I_a(s)/V_a(s)$ and $\Omega_m(s)/V_a(s)$. Build a Simulink block diagram by using these two transfer functions and find the armature current output $i_a(t)$ and the rotor speed $\omega_m(t)$.
- Choose θ_m and $\dot{\theta}_m$ as state variables and determine the state-space form of the system. Build a Simulink block diagram based on the state-space form and find the armature current output $i_a(t)$ and the rotor speed $\omega_m(t)$.
- Build a Simscape model of the DC motor and find the armature current output $i_a(t)$ and the rotor speed $\omega_m(t)$.

Solution

- This electromechanical system includes an armature circuit

$$R_a i_a = v_a - e_b$$

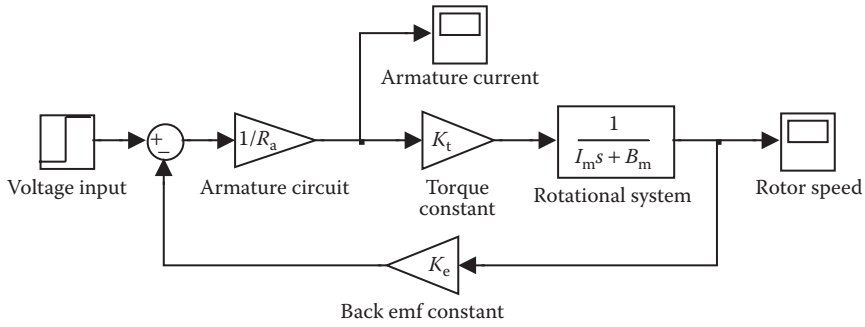
a rotational system

$$I_m \dot{\omega}_m + B_m \omega_m = \tau_m$$

and couplings between the electrical and mechanical subsystems

$$e_b = K_e \omega_m, \quad \tau_m = K_t i_a$$

Following [Figure 6.45](#), we can construct a Simulink block diagram ([Figure 6.67](#)), which shows the major components of the DC motor and their interconnections. The dynamics of the mechanical rotational system is represented using a **Transfer Fcn** block. The armature resistance, torque constant, and back emf constant, are represented using **Gain** blocks.

**FIGURE 6.67**

Simulink block diagram built based on the dynamics equations.

- b. Taking Laplace transform of the two equations given and denoting $s\Theta_m(s) = \Omega_m(s)$ yields

$$R_a I_a(s) + K_e \Omega(s) = V_a(s)$$

$$I_m s \Omega_m(s) + B_m \Omega_m(s) - K_t I_a(s) = 0$$

Using Cramer's rule to solve for the transfer functions $I_a(s)/V_a(s)$ and $\Omega_m(s)/V_a(s)$ yields

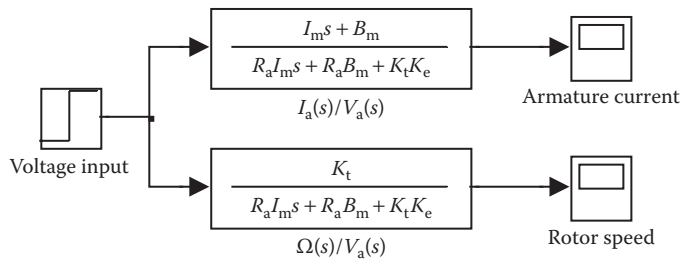
$$\frac{I_a(s)}{V_a(s)} = \frac{I_m s + B_m}{R_a I_m s + R_a B_m + K_t K_e}$$

$$\frac{\Omega_m(s)}{V_a(s)} = \frac{K_t}{R_a I_m s + R_a B_m + K_t K_e}$$

both of which can easily be represented using a Transfer Fcn block (Figure 6.68).

- c. Let $x_1 = \theta_m$, $x_2 = \dot{\theta}_m$, and $u = v_a$, and the state-variable equations are

$$\dot{x}_1 = \dot{\theta}_m = x_2$$

**FIGURE 6.68**

Simulink block diagram built based on the transfer functions.

$$\begin{aligned}\dot{x}_2 = \ddot{\theta}_m &= \frac{K_t}{I_m} i_a - \frac{B_m}{I_m} \dot{\theta}_m = \frac{K_t}{I_m} \left(\frac{v_a - K_e \dot{\theta}_m}{R_a} \right) - \frac{B_m}{I_m} \dot{\theta}_m \\ &= - \left(\frac{K_t K_e}{I_m R_a} + \frac{B_m}{I_m} \right) x_2 + \frac{K_t}{I_m R_a} u\end{aligned}$$

or in matrix form

$$\begin{Bmatrix} \dot{x}_1 \\ \dot{x}_2 \end{Bmatrix} = \begin{bmatrix} 0 & 1 \\ 0 & -\left(\frac{K_t K_e}{I_m R_a} + \frac{B_m}{I_m} \right) \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix} + \begin{bmatrix} 0 \\ \frac{K_t}{I_m R_a} \end{bmatrix} u$$

The output equations are

$$y_1 = i_a = \frac{v_a - K_e \dot{\theta}_m}{R_a} = -\frac{K_e}{R_a} x_2 + \frac{1}{R_a} u$$

$$y_2 = \dot{\theta}_m = x_2$$

or in matrix form

$$\begin{Bmatrix} y_1 \\ y_2 \end{Bmatrix} = \begin{bmatrix} 0 & -\frac{K_e}{R_a} \\ 0 & 1 \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix} + \begin{bmatrix} \frac{1}{R_a} \\ 0 \end{bmatrix} u$$

The system can be represented using a State-Space block (Figure 6.69) with A, B, C, and D matrices defined earlier. A Demux block from the Simulink library of Signal Routing is used to split the output vector signal into two scalar signals $i_a(t)$ and $\omega_m(t)$.

- d. The Simscape block diagram of the DC motor (Figure 6.70) consists of elements from two domains, electrical and mechanical rotational. Note that each domain requires at least one reference block. As shown in Figure 6.70,

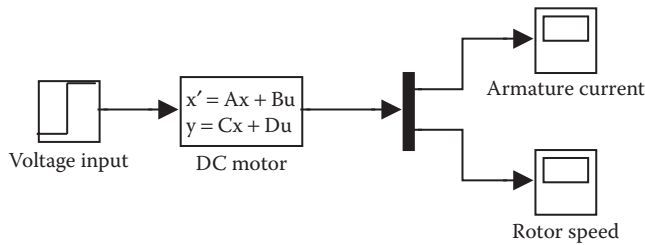


FIGURE 6.69

Simulink block diagram built based on the state-space form.

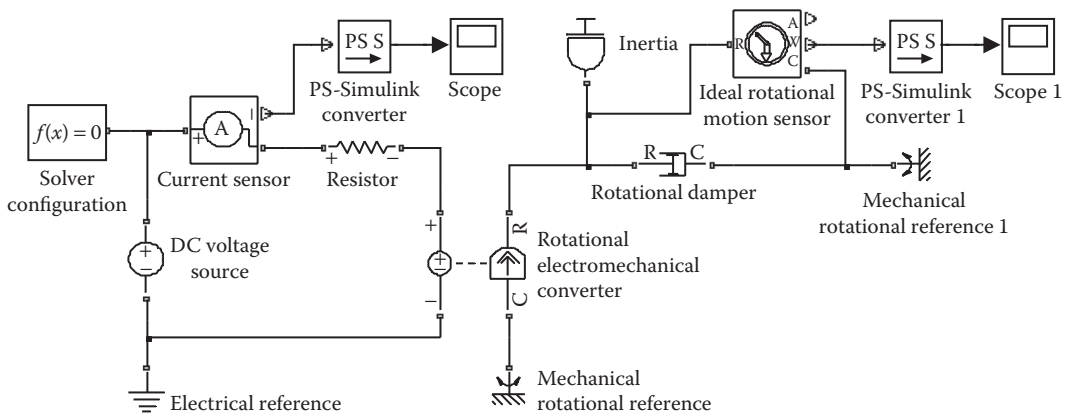


FIGURE 6.70
Simscape block diagram of the DC motor in Example 6.18.

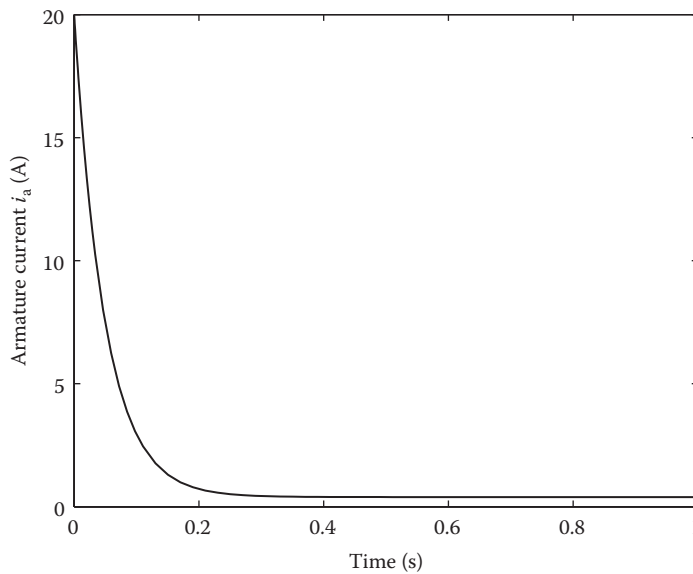
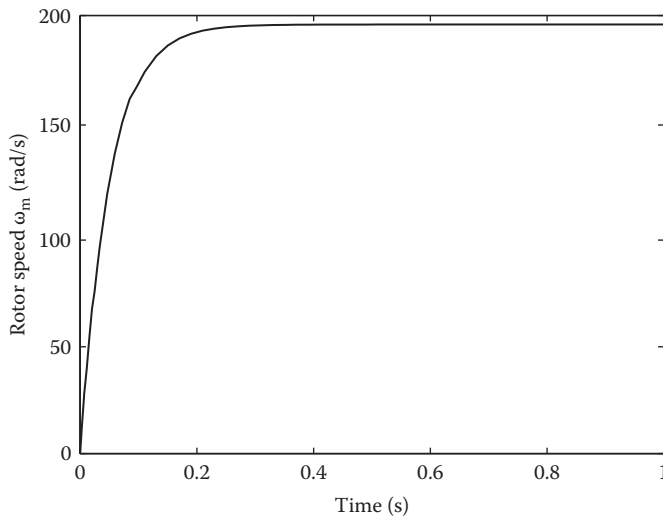


FIGURE 6.71
Armature current output $i_a(t)$ in Example 6.18.

both Electrical Reference and Mechanical Rotational Reference blocks are attached to the appropriate circuit.

Define all the parameters in MATLAB Workspace and run the Simulink or Simscape models. The result can be plotted as shown in [Figures 6.71](#) and [6.72](#).

**FIGURE 6.72**

Rotor speed output $\omega_m(t)$ in Example 6.18.

PROBLEM SET 6.6

1. Consider the RL circuit shown in [Figure 6.25](#) (Problem Set 6.2, Problem 2), in which $R = 35 \, \Omega$ and $L = 10 \, \text{H}$. When the switch is closed at 0 seconds, the circuit is driven by a 6 V DC voltage source. Assume that all initial conditions are zero.
 - a. Build a Simscape model of the physical system and find the loop current $i_L(t)$ and the voltage across the inductor $v_L(t)$.
 - b. Build a Simulink model of the system based on the differential equation relating i_L and v_a , and find the loop current $i_L(t)$.
 - c. Build a Simulink model of the system based on the transfer function $V_L(s)/V_a(s)$, and find the voltage across the inductor $v_L(t)$.
2. Consider the parallel RLC circuit shown in Example 6.2, in which $R = 2 \, \Omega$, $L = 1 \, \text{H}$, and $C = 0.5 \, \text{F}$. The circuit is driven by a controlled current source $i_a(t) = 10u(t)$, where $u(t)$ is a unit-step function.
 - a. Build a Simscape model of the physical system and find the voltage across the capacitor $v_C(t)$ and the current through the inductor $i_L(t)$.
 - b. Refer to the results obtained in Example 6.2. Build a Simulink model of the system based on the transfer function $V_C(s)/I_a(s)$ and find the voltage across the capacitor $v_C(t)$.
 - c. Refer to the results obtained in Example 6.2. Build a Simulink model of the system based on the transfer function $I_L(s)/I_a(s)$ and find the current through the inductor $i_L(t)$.
3. A simple low-pass filter can be realized by an RLC circuit (see [Figure 6.21](#) in Example 6.4), which passes signals with frequencies lower than a certain cutoff frequency and attenuates signals with a frequency higher than the cutoff frequency. Assume that the parameter values are $R = 500 \, \Omega$, $L = 100 \, \text{mH}$, and $C = 10 \, \mu\text{F}$.

The circuit is connected to an AC voltage source, which has an amplitude of 1 V and a varying frequency.

- a. Build a Simscape model of the physical system and find the output voltage $v_o(t)$ when the frequency of the input voltage is 500, 1000, and 2500 rad/s.
 - b. Build a Simulink model of the system based on the transfer function $V_o(s)/V_a(s)$, and verify the results obtained in Part (a).
4.  A simple band-pass filter can be realized by an RLC circuit ([Figure 6.73](#)), which passes frequencies within a certain range and attenuates frequencies outside that range. Assume that the parameter values are $R = 500 \Omega$, $L = 100 \text{ mH}$, and $C = 10 \mu\text{F}$. The circuit is connected to an AC voltage source, which has an amplitude of 1 V and a varying frequency.
- a. Build a Simscape model of the physical system and find the output voltage $v_o(t)$ when the frequency of the input voltage is 1000, 800, and 1200 rad/s.
 - b. Derive the transfer function $V_o(s)/V_a(s)$, build a Simulink model of the system based on this transfer function, and verify the results obtained in Part (a).
5.  Consider the op-amp integrator in [Figure 6.35](#). Assume that the parameter values are $R = 1 \text{ M}\Omega$ and $C = 1 \mu\text{F}$. Build a Simscape model of the op-amp circuit and find the output voltage $v_o(t)$ when the input voltage is $v_i = -0.1 \text{ V}$.
6.  Consider the op-amp circuit shown in [Figure 6.74](#), in which the parameter values are $C_1 = 0.8 \mu\text{F}$, $R_1 = 10 \text{ k}\Omega$, $C_2 = 80 \text{ pF}$, and $R_2 = 100 \text{ k}\Omega$. The circuit is connected to an AC voltage source, which has an amplitude of 1 V and a frequency of 200 Hz. Build a Simscape model of the op-amp circuit and find the output voltage $v_o(t)$.

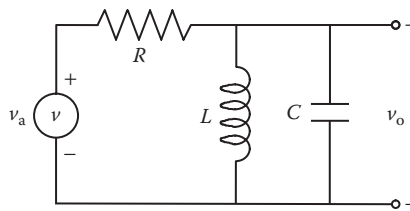


FIGURE 6.73
Problem 4.

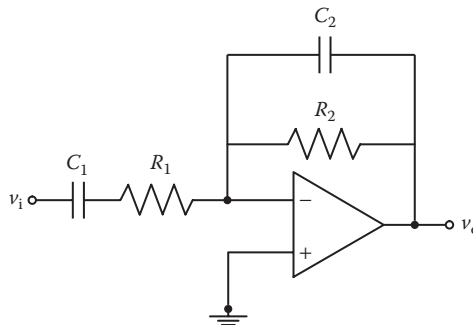


FIGURE 6.74
Problem 6.

6.7 Summary

This chapter was devoted to the modeling of electrical, electronic, and electromechanical systems. An electrical system or electrical circuit can be considered to be an interconnection of active and passive electrical elements. Active electrical elements include ideal current sources and ideal voltage sources, both of which can provide energy to the circuit and serve as the inputs. Passive electrical elements, including resistors, inductors, and capacitors, can either store or dissipate energy available in the circuit, but they cannot produce energy. The voltage–current relations for passive electrical elements are given as follows:

- Resistor: $v = Ri$, $i = \frac{v}{R}$
- Inductor: $v = L \frac{di}{dt}$, $i = \frac{1}{L} \int v dt$
- Capacitor: $v = \frac{1}{C} \int i dt$, $i = C \frac{dv}{dt}$

For the modeling of electrical, electronic, and electromechanical systems, Kirchhoff's voltage law and Kirchhoff's current law are the two main physical laws to derive the governing differential equations.

Kirchhoff's voltage law states that the algebraic sum of the voltages around a loop (closed path) must be zero,

$$\sum_j v_j = 0$$

where v_j is the voltage across the j th element in the loop.

Kirchhoff's current law states that the sum of the currents entering a node must be equal to the sum of the currents leaving that node. If we assign a positive sign to the current entering the node and a negative sign to the current leaving the node, the algebraic sum of the currents at the node must be zero,

$$\sum_j i_j = 0$$

where i_j is the current of the j th element at the node.

It is usually not easy to obtain a set of differential equations for complicated circuits. For this purpose, two systematic methods, the node method that relies on Kirchhoff's current law and the loop method that relies on Kirchhoff's voltage law, were introduced in [Section 6.2](#).

To represent a circuit model in state-space form, an appropriate set of state variables are normally chosen by identifying the energy storage elements. Both inductors and capacitors can store energy, and expressions for the stored electrical energy are given as follows:

- Inductor: $E(t) = \frac{1}{2} Li^2(t) - \frac{1}{2} Li^2(0)$
- Capacitor: $E(t) = \frac{1}{2} Cv^2(t) - \frac{1}{2} Cv^2(0)$

Generally, inductor currents and capacitor voltages are chosen as the state variables. To determine the state-space form of an electrical circuit, the expression of di_L/dt or dv_C/dt for each inductor or capacitor is needed. Based on the voltage–current relations, we have $di_L/dt = 1/Lv_L$ and $dv_C/dt = 1/Ci_C$. The problem is thus converted to expressing the inductor voltage v_L and the capacitor current i_C in terms of state variables and inputs by using Kirchhoff's laws and voltage–current relations for electrical elements.

For an op-amp, which is an electronic element used to amplify electrical signals and drive physical devices, the differential equation relating the output voltage and the input voltage can be derived by applying Kirchhoff's laws and the op-amp equation

$$v_+ \approx v_-$$

where v_+ and v_- are the voltages at the two input terminals of the op-amp.

For an electromechanical system, the dynamic model can be derived by applying electrical principles (e.g., Kirchhoff's laws) and mechanical principles (e.g., Newton's second law). The modeling of DC motors was discussed in [Section 6.4](#). For an armature-controlled motor, the torque produced by the motor is

$$\tau_m = K_t i_a$$

and the back emf generated in the armature due to the rotating motion is

$$e_b = K_e \omega = K_e \dot{\theta}$$

Armature-controlled motors are commonly used, in which a constant current source i_f is supplied to the field windings and the applied armature voltage v_a varies. Field-controlled motors are used in a different way, keeping the armature current i_a constant and letting the voltage v_f applied to the field windings vary. The torque generated by a field-controlled motor is proportional to the field current,

$$\tau_m = K_t i_f$$

The impedance concept provides an alternative to transfer functions and differential equations to obtain mathematical models of systems. The electrical impedance is defined as the ratio of the voltage to the current in the Laplace domain. The expressions of impedances for passive electrical elements are given as follows:

- Resistor: $Z(s) = R$
- Inductor: $Z(s) = Ls$
- Capacitor: $Z(s) = \frac{1}{Cs}$

Because the impedance can be viewed as a generalized resistance, it is easy to find the equivalent impedance for electrical elements connected in series or parallel and determine mathematical models of systems.

Review Problems

1. Determine the equivalent resistance R_{eq} for the circuit shown in Figure 6.75.
2. Find R_{13} and R_{32} for the voltage divider shown in Figure 6.76 so that the current is limited to 2 A when $v_i = 110$ V and $v_o = 100$ V.
3. Consider the second-order RC circuit shown in Figure 6.77. Assume that all the initial conditions are zero.
 - a. Use the node or loop method to derive the input–output differential equation relating v_o and v_a , and find the transfer function $V_o(s)/V_a(s)$.
 - b. Use the impedance method to determine the transfer function $V_o(s)/V_a(s)$, and compare with the result obtained Part (a).

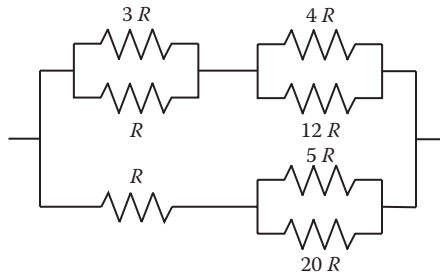


FIGURE 6.75
Problem 1.

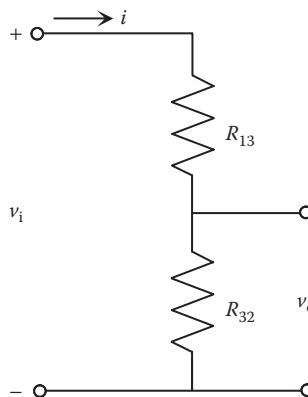
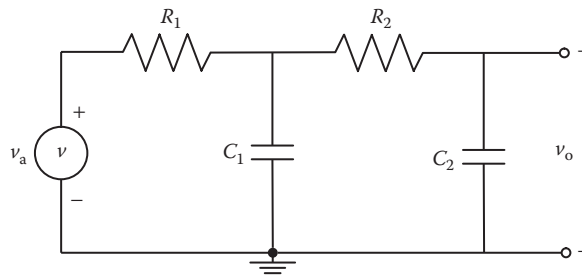
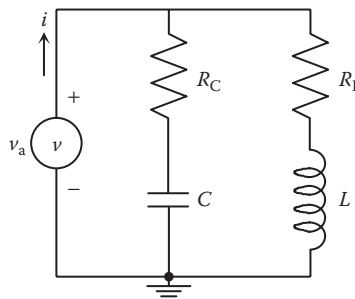


FIGURE 6.76
Problem 2.

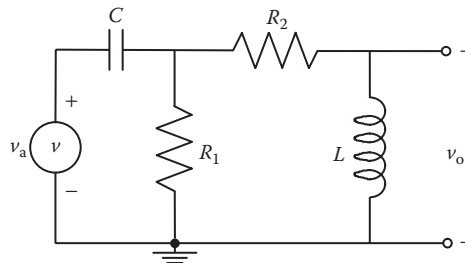
**FIGURE 6.77**

Problem 3.

4. Consider the RLC circuit shown in [Figure 6.78](#). Assume zero initial conditions.
 - a. Use the node or loop method to derive the input–output differential equation relating i and v_a , and find the transfer function $I(s)/V_a(s)$.
 - b. Use the impedance method to determine the transfer function $I(s)/V_a(s)$, and compare with the result obtained Part (a).
5. Consider the RLC circuit shown in [Figure 6.79](#), and assume zero initial conditions.
 - a. Use the node or loop method to derive the input–output differential equation relating v_o and v_a , and find the transfer function $V_o(s)/V_a(s)$.
 - b. Use the impedance method to determine the transfer function $V_o(s)/V_a(s)$, and compare with the result obtained Part (a).

**FIGURE 6.78**

Problem 4.

**FIGURE 6.79**

Problem 5.

6. Consider the circuit shown in [Figure 6.77](#) (Review Problems, Problem 3).
 - a. Determine a suitable set of state variables and obtain the state-space representation.
 - b. Find the transfer function directly from the state-space form and compare with the result obtained in Problem 3.
7. Repeat Problem 6 for the circuit shown in [Figure 6.78](#) (Review Problems, Problem 4).
8. Repeat Problem 7 for the circuit shown in [Figure 6.79](#) (Review Problems, Problem 5).
9. The op-amp circuit shown in [Figure 6.80](#) is an active low-pass filter. Derive the input–output differential equation relating the output voltage $v_o(t)$ and the input voltage $v_i(t)$. Assuming zero initial conditions, find the transfer function $V_o(s)/V_i(s)$ directly from the input–output equation.
10. Repeat Problem 9 for the op-amp circuit shown in [Figure 6.81](#), which represents an active band-pass filter.
11.  Consider the RLC circuit shown in [Figure 6.30](#) (Problem Set 6.2, Problem 15), where $R_1 = 100\ \Omega$, $L = 20\ \text{H}$, $R_2 = 400\ \Omega$, and $C = 1/120\ \text{F}$. The circuit is driven by a 100 V DC voltage source.
 - a. Build a Simscape model of the physical system and find the output voltage $v_o(t)$.
 - b. Build a Simulink model of the system based on the state-space form and find the output voltage $v_o(t)$.

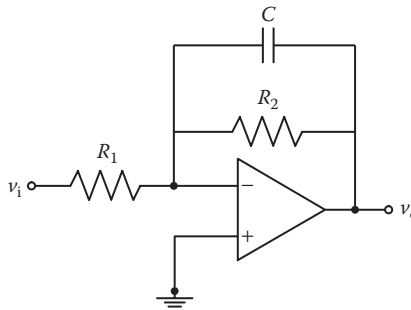


FIGURE 6.80
Problem 9.

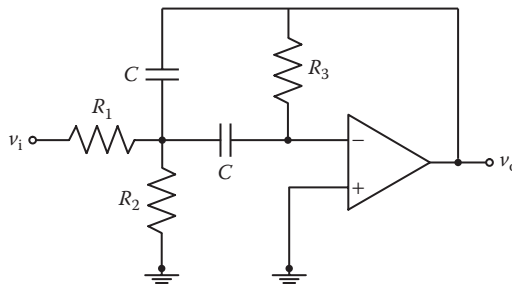
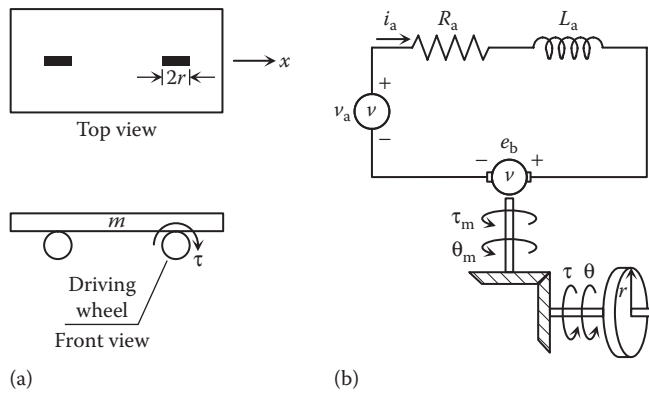


FIGURE 6.81
Problem 10.

**FIGURE 6.82**

Problem 12. (a) a DC motor-driven wheeled mobile robot and (b) the simplified drive system.

12. Case study

Consider the DC motor-driven wheeled mobile robot shown in Figure 6.82a, where m is the mass of the wheeled mobile robot, r is the radius of the driving wheel, and τ is the torque delivered to the wheeled mobile robot by the DC motor. For simplicity, the motion is restricted to one spatial dimension. Figure 6.82b shows the simplified drive system, including the equivalent electrical circuit of the DC motor, the rotor of the DC motor, the gears, and the driving wheel. The motor parameter values are armature inductance $L_a = 0.001$ H, resistance $R_a = 2.6 \Omega$, back emf constant $K_e = 0.008$ V s/rad, and torque constant $K_t = 0.008$ N m/A. The mass moment of inertia of the motor can be negligible. The gear ratio is $N = \theta/\theta_m = \tau_m/\tau = 1/3.7$. The wheel and axle mechanism converts the rotational motion to translation, and the wheel radius is $r = 0.00635$ m. The mass of the cart is $m = 0.455$ kg.

- Derive the equations of motion of the system.
- Choose the armature current i_a , the robot displacement x , and the robot velocity $\dot{x}$ as state variables, and find the state-space form of the system.
- Assuming zero initial conditions, find the transfer function $X(s)/V_a(s)$.
- Following Figure 6.45, build a Simulink block diagram by using the differential equations obtained in Part (a) and find the displacement output $x(t)$ when the voltage applied to the DC motor is a pulse function, $v_a(t) = 1$ V for $1 \leq t \leq 2$ s.
- Build a Simscape model of the wheeled mobile robot and find the displacement output $x(t)$ when the voltage applied to the DC motor is a pulse function, $v_a(t) = 1$ V for $1 \leq t \leq 2$ s.

Fluid and Thermal Systems

Fluid is a general term used to represent a gas or a liquid. A fluid is said to be incompressible if its density does not change with pressure. All gases are considered compressible, whereas liquids can be considered incompressible. Although real liquids are actually compressible, the changes in their densities are insignificant when pressure is varied. Fluid systems can be divided into pneumatics and hydraulics. A pneumatic system is one in which the fluid is compressible, whereas a hydraulic system is one in which the fluid is incompressible. A general type of incompressible liquid systems is the liquid-level system. These systems are operated by adjusting the heights or levels of liquids in storage tanks.

A thermal system is one in which thermal energy is stored or transferred. The mathematical model of a thermal system is often complicated because of the complex temperature distribution throughout the system. Partial, rather than ordinary, differential equations are required for precisely analyzing such a distributed-parameter system. This topic is beyond the scope of this text. To simplify analysis, a lumped-parameter model, rather than a distributed-parameter model, governed by ordinary differential equations, may be used to approximate the dynamics of the system.

The modeling of fluid and thermal systems is presented in this chapter. The three major systems that are discussed include pneumatic, liquid-level, and heat-transfer systems. For each, we first introduce the concepts of capacitance and resistance. It is useful to think of fluid and thermal systems as electrical circuits. Along with the basic elements, two main laws, the conservation of mass and the conservation of energy, are then used to develop mathematical models of fluid and thermal systems, respectively. The chapter concludes with simulation of fluid and thermal systems by using MATLAB®, Simulink®, and Simscape™ computer tools.

7.1 Pneumatic Systems

Pneumatic systems are often used in industry, particularly for pneumatic switches, pneumatic actuators, compressed-air engines, air brakes on buses and trucks, and so on. The working medium in a pneumatic system is compressible gas, typically air. To derive the mathematical model of a pneumatic system, it is important to understand the thermodynamic properties of gases.

7.1.1 Ideal Gases

The state of an amount of gas is determined by its pressure, volume, and temperature. In other words, pressure, volume, temperature, and mass are functionally related for gases. The ideal gas law is the model that is most often used to describe this relation. An ideal gas is a hypothetical gas for which the quantity pV/T is constant, where p is the absolute pressure of the gas, V is the volume, and T is the absolute temperature (K or °R). All real gases

behave as ideal gases if the pressure is low enough and the temperature is high enough. At low pressure and moderate temperature, real gases may be approximated as ideal gases to simplify calculations.

The ideal gas law states the relationship

$$pV = nR_u T \quad (7.1)$$

where:

n is the number of moles of the gas

R_u is the universal gas constant

The mole is the unit of the amount of substance. A mole of an element or a compound contains 6.023×10^{23} atoms or molecules. The numerical value of the universal gas constant R_u is $8314.3 \text{ N}\cdot\text{m}/(\text{kg}\cdot\text{mol}\cdot\text{K})$ or $1545.3 \text{ ft}\cdot\text{lb}/(\text{lb}\cdot\text{mol}\cdot^\circ\text{R})$.

The amount of substance can also be given in mass, instead of moles. The number of moles n is equal to m/M , where m is the mass and M is the molar mass. Therefore, an alternative form of the ideal gas law is

$$pV = mR_g T \quad (7.2)$$

where $R_g = R_u/M$ is the specific gas constant that depends on the particular type of gas. For dry air, $R_g = 287.06 \text{ N}\cdot\text{m}/(\text{kg}\cdot\text{K})$ or $1716.6 \text{ ft}\cdot\text{lb}/(\text{slug}\cdot^\circ\text{R})$. The ideal gas law can be used to solve for one of the four variables (p , V , T , and m) if the other three are known.

For a particular thermodynamic process from state 1 to state 2, the ideal gas equation can be simplified. Assume that the mass is constant. The following is a list of five possible processes, in which the state number is denoted by the subscript.

1. *Isobaric (or constant-pressure) process* ($p_1 = p_2$): The ideal gas law implies that $T_1/V_1 = T_2/V_2$ or $T_1/T_2 = V_1/V_2$. When heat is added to the gas, some of it increases the temperature and some expands the volume to exert external work.
2. *Isochoric (or constant-volume) process* ($V_1 = V_2$): The ideal gas law implies that $T_1/p_1 = T_2/p_2$ or $T_1/T_2 = p_1/p_2$. Because the volume is constant, no external work is done. The heat added to the gas only increases the temperature.
3. *Isothermal (or constant-temperature) process* ($T_1 = T_2$): The ideal gas law implies that $p_1V_1 = p_2V_2$ or $p_1/p_2 = V_2/V_1$. The heat added to the gas only does external work.
4. *Isentropic process (reversible adiabatic process)*: An adiabatic process is a process in which no heat is transferred to or from the gas. A reversible process is a process that, after it has taken place, can be reversed, and it causes no change in the thermodynamic conditions of either the system or its surroundings. Any reversible adiabatic process is an isentropic process. This process is described by $p_1V_1^\gamma = p_2V_2^\gamma$, where γ is defined as the heat capacity ratio.
5. *Polytropic process*: It is the most general thermodynamic process. The process is described by

$$p \left(\frac{V}{m} \right)^n = \frac{p}{\rho^n} = \text{constant} \quad (7.3)$$

where ρ is the density of the gas. For an ideal gas with a constant mass, this process reduces to the previous four processes if n is chosen as 0, ∞ , 1, or γ , respectively.

7.1.2 Pneumatic Capacitance

Fluid capacitance is the relation between the stored fluid mass and the resulting pressure caused by the stored mass. Specifically, the pneumatic capacitance C is defined as the ratio of the change in stored gas mass to the change in gas pressure:

$$C = \frac{dm}{dp} \quad (7.4)$$

For a container of constant volume V , with a gas of density ρ , Equation 7.4 can be rewritten as

$$C = \frac{d(\rho V)}{dp} = V \frac{d\rho}{dp} \quad (7.5)$$

For a polytropic process, we have

$$\frac{dp}{d\rho} = n\rho^{n-1} \left(\frac{p}{\rho^n} \right) = \frac{np}{\rho} \quad (7.6)$$

Introducing the ideal gas law presented in Equation 7.2 gives

$$\frac{p}{\rho} = \frac{pV}{m} = R_g T \quad (7.7)$$

Substituting it into Equation 7.6 gives

$$\frac{d\rho}{dp} = \frac{1}{nR_g T} \quad (7.8)$$

Thus, the capacitance of the container is

$$C = \frac{V}{nR_g T} \quad (7.9)$$

Example 7.1: Pneumatic Capacitance

Dry air passes through a valve into a rigid 27 m³ container at a constant temperature of 25°C (298 K). The process is assumed to be isothermal. Determine the capacitance of the air container.

Solution

The filling of the container is modeled as an isothermal process. In Equation 7.9, let $n = 1$, and we have

$$C = \frac{V}{R_g T} = \frac{27}{287.06 \times 298} = 3.16 \times 10^{-4} \text{ kg} \cdot \text{m}^2/\text{N}$$

Note that the same container can have a different pneumatic capacitance. The value of C depends on the type of gas, the temperature of gas, and the type of thermodynamic process.

7.1.3 Modeling of Pneumatic Systems

It is rather difficult to precisely model pneumatic systems, owing to their highly nonlinear dynamics. To simplify the modeling, each mass storage element in a pneumatic system can be represented by a capacitance element, and the resistance due to a valve, an orifice, or pipe wall friction can be represented by a resistance element. Such a simple model may be adequate to describe the dynamic behavior of the real system.

Consider a pneumatic system shown in [Figure 7.1](#), where p_i is the inlet pressure, q_i is the volume flow rate (m^3/s), and p and ρ are the pressure and the density of the gas, respectively, in a container of constant volume V . The gas passes through a valve and flows into the rigid container by the pressure difference, $\Delta p = p_i - p$. Note that the gas meets resistance when flowing through the valve. The valve resistance depends on the pressure p and the mass flow rate q_m (kg/s). Generally, the p versus q_m curve is nonlinear. Thus, the value of the valve resistance R , which is defined as the slope of the curve, varies with the operating conditions. The definition of the resistance R is valid for both pneumatic and hydraulic systems, and more details will be given in [Section 7.2](#).

Although the valve resistance R is nonlinear, it may be linearized about an operating point. For a constant pressure p_i at the inlet of the valve and a constant volume flow rate q_i through the valve, the resistance R is given as

$$R = \frac{\Delta p}{\Delta q_m} = \frac{p_i - p}{\rho_i q_i} \quad (7.10)$$

Combining with the capacitance of the container given in Equation 7.9, the pneumatic system can be represented using a resistance–capacitance model.

The differential equation of the system can be derived by applying the law of conservation of mass:

$$\frac{dm}{dt} = q_{mi} - q_{mo} \quad (7.11)$$

That is, the rate of mass increase in the container equals the mass flow rate into the container minus the mass flow rate out of the container. Note that

$$\frac{dm}{dt} = \frac{dm}{dp} \frac{dp}{dt} = C \frac{dp}{dt} \quad (7.12)$$

and

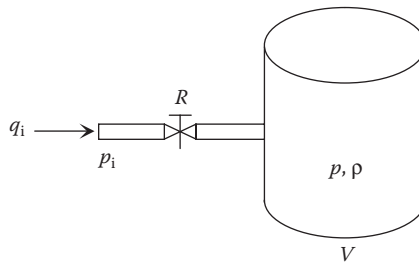


FIGURE 7.1

A pneumatic system with gas flowing into a container of constant volume.

$$q_{mi} = \rho_i q_i = \frac{p_i - p}{R} \quad (7.13)$$

For the pneumatic system in [Figure 7.1](#), the mass flow rate out of the container is $q_{mo} = 0$. Thus, Equation 7.11 can be rewritten as

$$C \frac{dp}{dt} = \frac{p_i - p}{R} \quad (7.14)$$

or

$$RC \frac{dp}{dt} + p = p_i \quad (7.15)$$

which is a first-order ordinary differential equation of the pressure p . Introducing the expression of the capacitance C given by Equation 7.9, we find

$$\frac{RV}{nR_g T} \frac{dp}{dt} + p = p_i \quad (7.16)$$

Either one of Equations 7.15 and 7.16 is the mathematical model of the pneumatic system undergoing a polytropic process with an ideal gas. It is also valid for an isobaric, isochoric, isothermal, or isentropic process if the value of n in the capacitance C is chosen as 0, ∞ , 1, or γ , respectively.

Example 7.2: A Pneumatic System

Dry air at a constant temperature of 20°C (293 K) passes through a valve into a rigid cubic container of 1 m on each side ([Figure 7.2](#)). The pressure p_i at the inlet of the valve is constant, and it is greater than p . The valve resistance is approximately linear, and $R = 1000$ Pa·s/kg. Assume that the filling process is isothermal. Develop a mathematical model of the pressure p in the container.

Solution

Applying the law of conservation of mass gives

$$\frac{dm}{dt} = \rho_i q_i$$

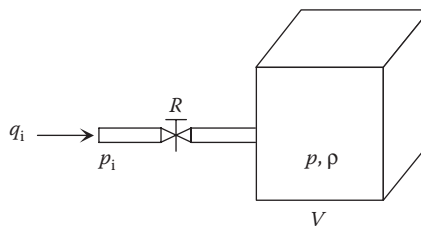


FIGURE 7.2

A pneumatic system with a rigid cubic container.

Note that

$$\frac{dm}{dt} = \frac{dm}{dp} \frac{dp}{dt} = C \frac{dp}{dt}$$

Air at room temperature and low pressure can be approximated as an ideal gas. For an isothermal process, the pneumatic capacitance of the container is

$$C = \frac{V}{R_{\text{air}} T}$$

The linear valve resistance R is defined as

$$R = \frac{p_i - p}{\rho_i q_i}$$

Thus, the differential equation of the system is

$$\frac{V}{R_{\text{air}} T} \frac{dp}{dt} = \frac{p_i - p}{R}$$

or

$$\frac{RV}{R_{\text{air}} T} \frac{dp}{dt} + p = p_i$$

where $RV/(R_{\text{air}} T) = 1000 \times 1^3 / (287.06 \times 293) = 1.19 \times 10^{-2} \text{ s}$.

PROBLEM SET 7.1

1. A car has an internal volume of 2.8 m^3 . If the sun heats the car from a temperature of 20°C to a temperature of 35°C , what will be the pressure inside the car? Assume the pressure is initially 1 atm.
2. Find the pneumatic capacitance of dry air in a rigid container with volume 15 ft^3 for an isothermal process. Assume that the air is initially at ambient temperature of 76°F .
3. Figure 7.3 shows a pneumatic system, in which the pneumatic capacitances of the two rigid containers are C_1 and C_2 , respectively. Dry air at a constant temperature passes through a valve of linear resistance R_1 into container 1. The pressure p_i at the inlet of the valve is constant, and it is greater than p_1 . The air flows from container 1 to container 2 through a valve of linear resistance R_2 . Derive the differential equations in terms of the pressures p_1 and p_2 . Write the equations in second-order matrix form.

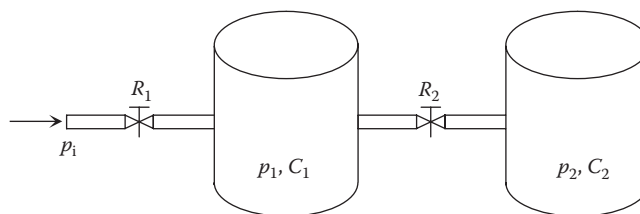


FIGURE 7.3
Problem 3.

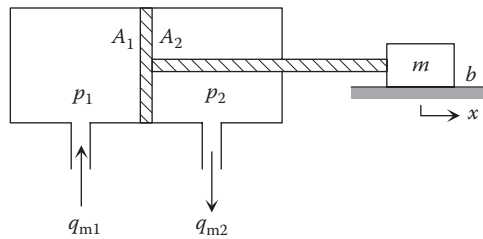


FIGURE 7.4
Problem 4.

4. [Figure 7.4](#) shows a pneumatic piston, which can exert a force in one direction and serve as a translational actuator. The displacement of the piston is x , the total mass of the piston and the load is m , and the damping coefficient is b . Use subscript i to denote the chamber index and $i = 1, 2$. As shown in [Figure 7.4](#), p_i is the absolute pressure in the chamber, q_{mi} is the mass flow rate at the port, and A_i is the effective area of the piston. The two chambers have the same initial volume of V_0 . Assuming ideal gas and isothermal process, R is the specific gas constant and T is the absolute gas temperature. Derive the dynamic equations of the pressure change in the pneumatic chambers and the equation of motion of the mass block.

7.2 Liquid-Level Systems

Unlike gases, most liquids are generally considered incompressible, and this approximation greatly simplifies the modeling of hydraulic systems. A general category of hydraulic systems is liquid-level systems, which often appear in water treatment, water supply, and other chemical processing applications. Such a system usually consists of storage tanks interconnected to other systems through valves, pumps, or pistons.

The dynamic behavior of a liquid-level system can be described using volume flow rate q , pressure p , and liquid height h . Note that the hydrostatic pressure, rather than the dynamic pressure, will be used in the modeling of liquid-level systems. The hydrostatic pressure is defined as the pressure that exists in a fluid at rest. It is caused by the weight of the fluid. For a liquid of density ρ , the absolute pressure p and the liquid height h are related by

$$p = p_a + \rho gh \quad (7.17)$$

where p_a is the atmospheric pressure.

7.2.1 Hydraulic Capacitance

As introduced in [Section 7.1](#), fluid capacitance is the ratio of the change in stored mass to the change in pressure. Because the density of an incompressible liquid is constant, the change in mass is equivalent to the change in volume. Some books define the capacitance for hydraulic systems in terms of volume instead of mass,

$$C_V = \frac{dV}{dp} \quad (7.18)$$

which is related to Equation 7.4 as $C = \rho C_V$. The definition given in Equation 7.4 is used for both pneumatic and hydraulic systems in this text.

To find the expression of capacitance for a hydraulic system, let us consider a container whose cross-sectional area varies with the liquid height. The mass stored in the container can be determined by integrating ρA from the base of the container to the top of the liquid,

$$m = \int_0^h \rho A(y) dy \quad (7.19)$$

By Equation 7.4,

$$C = \frac{dm}{dp} = \frac{dm}{dh} \frac{dh}{dp} \quad (7.20)$$

Note that Equation 7.19 implies that

$$\frac{dm}{dh} = \rho A(h) \quad (7.21)$$

Also, Equation 7.17 gives

$$\frac{dp}{dh} = \frac{d}{dh}(p_a + \rho gh) = \rho g \quad (7.22)$$

Thus, the hydraulic capacitance of the container is

$$C = \rho A(h) \frac{1}{\rho g} = \frac{A(h)}{g} \quad (7.23)$$

If the cross-sectional area of the container is constant, then we have $C = A/g$. Unlike the pneumatic capacitance (see Equation 7.9), which depends on the type of gas and its temperature, the hydraulic capacitance does not depend on any liquid properties.

Example 7.3: Hydraulic Capacitance of a Conical Tank

Derive the capacitance of the conical tank shown in Figure 7.5a by using

- $C = dm/dp$.
- $C = A(h)/g$.

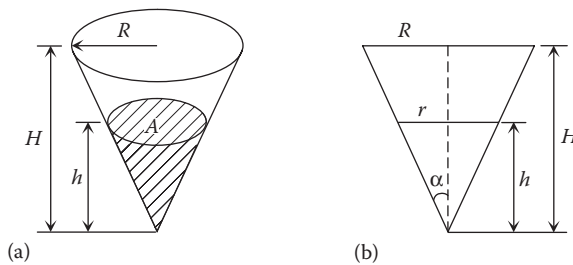


FIGURE 7.5

A conical tank: (a) three-dimensional view and (b) cross-sectional view.

Solution

- a. From Figure 7.5b, the radius r of the cross-section A at an arbitrary height is

$$r = h \tan \alpha = h \frac{R}{H}$$

Thus, the volume of the liquid is

$$V(h) = \frac{1}{3} \pi r^2 h = \frac{1}{3} \frac{\pi R^2}{H^2} h^3$$

and the stored mass is

$$m = \rho V(h) = \frac{1}{3} \frac{\rho \pi R^2}{H^2} h^3$$

Note that the pressure caused by the height of the liquid is

$$p = p_a + \rho g h$$

which gives

$$\frac{dp}{dh} = \rho g$$

Thus, the capacitance of the conical tank is

$$C = \frac{dm}{dp} = \frac{dm}{dh} \frac{dh}{dp} = \left(\frac{\rho \pi R^2}{H^2} h^2 \right) \left(\frac{1}{\rho g} \right) = \frac{\pi R^2 h^2}{H^2 g}$$

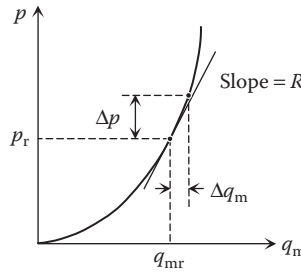
- b. The hydraulic capacitance can also be derived directly by using $C = A(h)/g$, which yields

$$C = \frac{\pi r^2}{g} = \left(\frac{\pi}{g} \right) \left(\frac{R^2 h^2}{H^2} \right) = \frac{\pi R^2 h^2}{H^2 g}$$

7.2.2 Hydraulic Resistance

When liquid flows through a pipe, a valve, or an orifice, the liquid meets resistance, which creates a reduction in the pressure of the liquid. The pressure difference is associated with the mass flow rate q_m in a nonlinear relationship, $p = f(q_m)$, as illustrated in Figure 7.6. The slope of the curve is defined as the hydraulic resistance R , which depends on the reference mass flow rate q_{mr} and the reference pressure p_r . The expression of the hydraulic resistance R is given by

$$R = \left. \frac{dp}{dq_m} \right|_{(q_{mr}, p_r)} \quad (7.24)$$

**FIGURE 7.6**

Linearized resistance near a reference point.

which is also used for pneumatic systems. Near a reference operating point (q_{mr}, p_r) , we can perform linearization and obtain the linearized resistance, which is

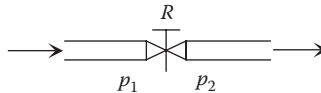
$$R = \frac{\Delta p}{\Delta q_m} = \frac{p - p_r}{q_m - q_{mr}} \quad (7.25)$$

The resistance due to a valve, an orifice, or pipe wall friction can be represented by the valve-like symbol, as in Figure 7.7. In many fluid systems, multiple valves, orifices, or pipes are used. They are arranged in different ways, such as in series or parallel connections. The equivalent linear hydraulic resistances can be obtained similar to electrical resistances.

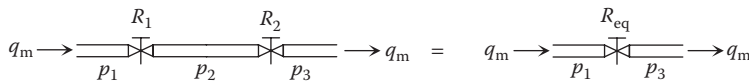
Figure 7.8 shows resistances in series. Note that the mass flow rate remains the same through each resistance. The pressure reductions across the resistances R_1 and R_2 are $p_1 - p_2 = R_1 q_m$ and $p_2 - p_3 = R_2 q_m$, respectively. Consequently, the total pressure reduction across the two resistances in series is $p_1 - p_3 = (R_1 + R_2) q_m$. Comparing this result with the equivalent fluid system, $p_1 - p_3 = R_{eq} q_m$, we have

$$R_{eq} = R_1 + R_2 \quad (7.26)$$

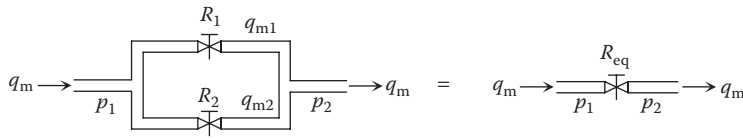
A fluid system with resistances in parallel is shown in Figure 7.9. Note that the pressure reduction across each resistance must be the same. The mass flow rates through the resistances R_1 and R_2 are $q_{m1} = (p_1 - p_2)/R_1$ and $q_{m2} = (p_1 - p_2)/R_2$, respectively. Therefore, the total

**FIGURE 7.7**

A symbol for hydraulic resistance.

**FIGURE 7.8**

Equivalence for series hydraulic resistances.

**FIGURE 7.9**

Equivalence for parallel hydraulic resistances.

mass flow rate through the two resistances in parallel is $q_{m1} + q_{m2} = (p_1 - p_2)(1/R_1 + 1/R_2)$. Comparing this result with the equivalent fluid system, $q_m = (p_1 - p_2)/R_{eq}$, we have

$$\frac{1}{R_{eq}} = \frac{1}{R_1} + \frac{1}{R_2} \quad (7.27)$$

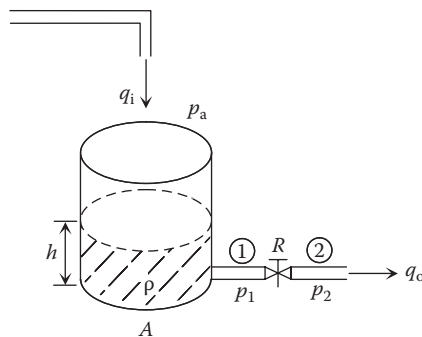
7.2.3 Modeling of Liquid-Level Systems

To obtain a simple model of a liquid-level system, we will use a capacitance element to represent each storage tank and a resistance element to represent each valve in the system. The resulting mathematical model may adequately describe the dynamics of the real system. As in the modeling of pneumatic systems, the basic law used to derive the differential equation of a liquid-level system is the law of conservation of mass, presented in Equation 7.11: $dm/dt = q_{mi} - q_{mo}$. That is, the time rate of change of fluid mass in a tank equals the mass flow rate into the tank minus the mass flow rate out of the tank.

Consider a single tank with a valve, as shown in Figure 7.10, where p_a is the atmospheric pressure and q_i and q_o are the volume flow rates into and out of the tank, respectively. The cross-sectional area of the tank is A , and the liquid height is h . The liquid leaves the tank through the valve, for which the hydraulic resistance is R . The density of the liquid is constant, ρ .

Next, we will show how to derive the differential equation of the system by applying the law of conservation of mass, presented in Equation 7.11. Note that the total fluid mass in the tank is ρAh . For constant cross-sectional area and constant density, the left-hand side of Equation 7.11 can be rewritten as

$$\frac{dm}{dt} = \frac{d}{dt}(\rho Ah) = \rho A \frac{dh}{dt} \quad (7.28)$$

**FIGURE 7.10**

A single-tank liquid-level system with a valve.

The right-hand side of Equation 7.11 can be rewritten as

$$q_{mi} - q_{mo} = \rho q_i - \rho q_o \quad (7.29)$$

Labeling point 1 at the upstream side of the valve and point 2 at the downstream side of the valve, the hydraulic resistance of the valve can be expressed as

$$R = \frac{\Delta p}{\Delta q_m} = \frac{p_1 - p_2}{\rho q_o} \quad (7.30)$$

Assume that the pressure p_1 can be approximated as the hydrostatic pressure, $p_1 = p_a + \rho gh$, and the pressure p_2 can be approximated as the atmospheric pressure, $p_2 = p_a$. Substituting p_1 and p_2 into Equation 7.30 gives

$$R = \frac{\rho gh}{\rho q_o} \quad (7.31)$$

Consequently,

$$\rho q_i - \rho q_o = \rho q_i - \frac{\rho gh}{R} \quad (7.32)$$

or

$$q_{mi} - q_{mo} = \rho q_i - \frac{\rho gh}{R} \quad (7.33)$$

Substituting Equations 7.28 and 7.33 into Equation 7.11 results in

$$\rho A \frac{dh}{dt} = \rho q_i - \frac{\rho gh}{R} \quad (7.34)$$

Rearranging the equation gives

$$\frac{RA}{g} \frac{dh}{dt} + h = \frac{R}{g} q_i \quad (7.35)$$

which is a first-order ordinary differential equation relating the liquid height h and the inlet volume flow rate q_i . Introducing the expression of the hydraulic capacitance C , given by Equation 7.23, we can rewrite Equation 7.35 as

$$RC \frac{dh}{dt} + h = \frac{R}{g} q_i \quad (7.36)$$

Equation 7.36 describes the dynamic behavior of a single-tank liquid-level system with capacitance C and resistance R , as shown schematically in [Figure 7.10](#).

Hydraulic systems are usually connected with pumps, which can be considered as pressure sources. The following example shows how to derive the governing differential equation for a single-tank liquid-level system with a pump.

Example 7.4: A Single-Tank Liquid-Level System with a Pump

Consider the single-tank liquid-level system shown in Figure 7.11, where a pump is connected to the bottom of the tank through a valve of linear resistance R . The inlet to the pump is open to the atmosphere, and the pressure of the fluid increases by Δp when crossing the pump. Derive the differential equation relating the liquid height h and the volume flow rate q_o at the outlet. The tank's cross-sectional area A is constant. The density ρ of the liquid is constant.

Solution

We begin by applying the law of conservation of mass to the tank,

$$\frac{dm}{dt} = q_{mi} - q_{mo}$$

The fluid mass inside the tank is ρAh . For constant fluid density and constant cross-sectional area,

$$\frac{dm}{dt} = \rho A \frac{dh}{dt}$$

The mass flow rate into the tank is

$$q_{mi} = \frac{p_1 - p_2}{R}$$

where $p_1 = p_a + \Delta p$ and $p_2 = p_a + \rho gh$, which is equal to the hydrostatic pressure at the bottom of the tank. Thus,

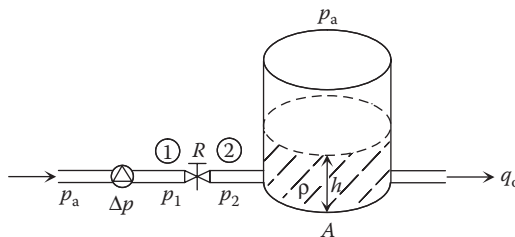
$$q_{mi} = \frac{\Delta p - \rho gh}{R}$$

The mass flow rate out of the tank can be expressed in terms of the volume flow rate q_o , as

$$q_{mo} = \rho q_o$$

Substituting these expressions into the law of conservation of mass gives

$$\rho A \frac{dh}{dt} = \frac{\Delta p - \rho gh}{R} - \rho q_o$$

**FIGURE 7.11**

A single-tank liquid-level system with a pump.

Rearranging the equation gives

$$\rho A \frac{dh}{dt} + \frac{\rho g}{R} h - \frac{\Delta p}{R} = -\rho q_o$$

For a liquid-level system with two or more tanks, we apply the law of conservation of mass to each tank.

Example 7.5: A Two-Tank Liquid-Level System

Figure 7.12 shows a liquid-level system in which two tanks have cross-sectional areas A_1 and A_2 , respectively. A pump is connected to the bottom of tank 1 through a valve of linear resistance R_1 . The liquid flows from tank 1 to tank 2 through a valve of linear resistance R_2 and leaves tank 2 through a valve of linear resistance R_3 . The density ρ of the liquid is constant.

- Derive the differential equations in terms of the liquid heights h_1 and h_2 . Write the equations in second-order matrix form.
- Assume the pump pressure Δp as the input and the liquid heights h_1 and h_2 as the outputs. Determine the state-space form of the system.

Solution

- Applying the law of conservation of mass to tank 1 gives

$$\frac{dm}{dt} = q_{mi} - q_{mo}$$

where

$$\frac{dm}{dt} = \rho A_1 \frac{dh_1}{dt}$$

The mass flow rate entering and leaving tank 1 can be written as

$$q_{mi} = \frac{(p_a + \Delta p) - (p_a + \rho g h_1)}{R_1} = \frac{\Delta p - \rho g h_1}{R_1}$$

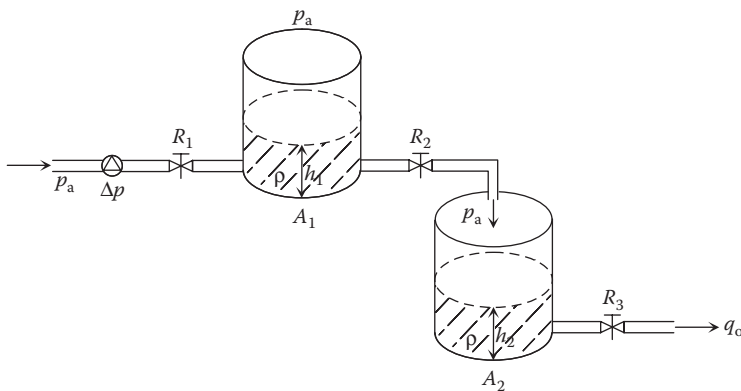


FIGURE 7.12

A two-tank liquid-level system.

and

$$q_{mo} = \frac{(p_a + \rho g h_1) - p_a}{R_2} = \frac{\rho g h_1}{R_2}$$

Substituting these expressions results in

$$\rho A_1 \frac{dh_1}{dt} = \frac{\Delta p - \rho g h_1}{R_1} - \frac{\rho g h_1}{R_2}$$

which can be rearranged as

$$\rho A_1 \frac{dh_1}{dt} + \rho g h_1 \left(\frac{1}{R_1} + \frac{1}{R_2} \right) = \frac{\Delta p}{R_1}$$

Applying the law of conservation of mass to tank 2 gives

$$\frac{dm}{dt} = q_{mi} - q_{mo}$$

where

$$\frac{dm}{dt} = \rho A_2 \frac{dh_2}{dt}$$

The mass flow rate entering and leaving tank 2 can be written as

$$q_{mi} = \frac{(p_a + \rho g h_1) - p_a}{R_2} = \frac{\rho g h_1}{R_2}$$

and

$$q_{mo} = \frac{(p_a + \rho g h_2) - p_a}{R_3} = \frac{\rho g h_2}{R_3}$$

Substituting these expressions results in

$$\rho A_2 \frac{dh_2}{dt} = \frac{\rho g h_1}{R_2} - \frac{\rho g h_2}{R_3}$$

which can be rearranged as

$$\rho A_2 \frac{dh_2}{dt} - \frac{\rho g h_1}{R_2} + \frac{\rho g h_2}{R_3} = 0$$

The system of differential equations in second-order matrix form is found to be

$$\begin{bmatrix} \rho A_1 & 0 \\ 0 & \rho A_2 \end{bmatrix} \begin{Bmatrix} \frac{dh_1}{dt} \\ \frac{dh_2}{dt} \end{Bmatrix} + \begin{bmatrix} \frac{\rho g}{R_1} + \frac{\rho g}{R_2} & 0 \\ -\frac{\rho g}{R_2} & \frac{\rho g}{R_3} \end{bmatrix} \begin{Bmatrix} h_1 \\ h_2 \end{Bmatrix} = \begin{Bmatrix} \frac{\Delta p}{R_1} \\ 0 \end{Bmatrix}$$

b. As specified, the state, the input, and the output are

$$\mathbf{x} = \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix} = \begin{Bmatrix} h_1 \\ h_2 \end{Bmatrix}, \quad u = \Delta p, \quad \mathbf{y} = \begin{Bmatrix} h_1 \\ h_2 \end{Bmatrix}$$

The state-variable equations are

$$\dot{x}_1 = \frac{dh_1}{dt} = -\frac{g}{A_1} \left(\frac{1}{R_1} + \frac{1}{R_2} \right) h_1 + \frac{1}{\rho A_1 R_1} \Delta p$$

$$\dot{x}_2 = \frac{dh_2}{dt} = \frac{g}{A_2 R_2} h_1 - \frac{g}{A_2 R_3} h_2$$

or in matrix form

$$\begin{Bmatrix} \dot{x}_1 \\ \dot{x}_2 \end{Bmatrix} = \begin{bmatrix} -\frac{g}{A_1} \left(\frac{1}{R_1} + \frac{1}{R_2} \right) & 0 \\ \frac{g}{A_2 R_2} & -\frac{g}{A_2 R_3} \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix} + \begin{bmatrix} \frac{1}{\rho A_1 R_1} \\ 0 \end{bmatrix} u$$

The output equation is

$$\begin{Bmatrix} y_1 \\ y_2 \end{Bmatrix} = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix} + \begin{bmatrix} 0 \\ 0 \end{bmatrix} u$$

PROBLEM SET 7.2

1. Derive the capacitance of the tank shown in [Figure 7.13](#). There is an opening at the top of the tank at height H .
2. Derive the capacitance of the tank shown in [Figure 7.14](#).
3. Consider the rectangular tank in [Figure 7.15a](#) and the pyramid tank in [Figure 7.15b](#). The volume flow rate into each tank through a pipe is q_i . The liquid leaves each tank through a valve of linear resistance R . The density ρ of the liquid is constant.
 - a. Derive the dynamic model of the liquid height h for each tank.
 - b. For each tank, write the differential equation in terms of hydraulic capacitance and hydraulic resistance. Compare the models for the two single-tank systems.

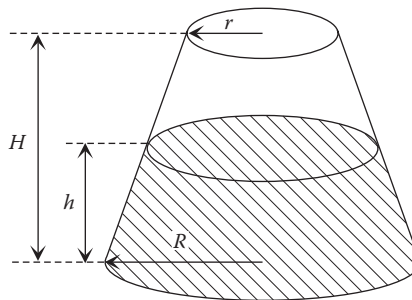


FIGURE 7.13
Problem 1.

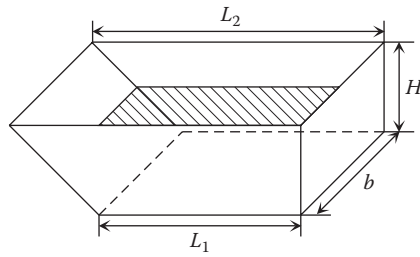


FIGURE 7.14
Problem 2.

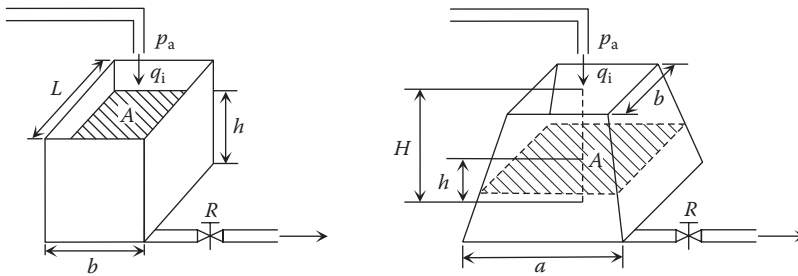


FIGURE 7.15
Problem 3.

4. Consider the single-tank liquid-level system shown in [Figure 7.16](#), where the volume flow rate into the tank through a pipe is q_i . A pump is connected to the bottom of the tank through a valve of linear resistance R_1 . The pressure of the fluid increases by Δp when crossing the pump. The liquid leaves the tank through a valve of linear resistance R_2 . Derive the differential equation relating the liquid height h and the volume flow rate q_i at the inlet. The tank's cross-sectional area is constant. The density ρ of the liquid is constant.
5. Consider the two-tank liquid-level system shown in [Figure 7.17](#). The liquid is pumped into tanks 1 and 2 through valves of linear resistances R_1 and R_2 , respectively. The pressure of the fluid increases by Δp when crossing the pump. The cross-sectional areas of the two tanks are A_1 and A_2 , respectively. The liquid flows from tank 1 to tank 2 through a valve of linear resistance R_3 and leaves tank 2

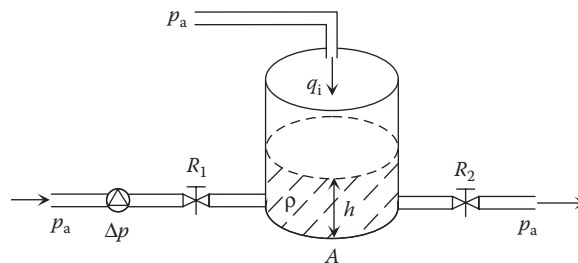


FIGURE 7.16
Problem 4.

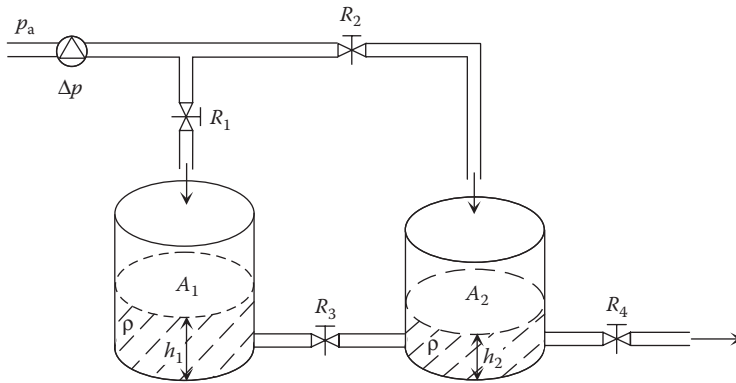


FIGURE 7.17
Problem 5.

through a valve of linear resistance R_4 . The density ρ of the liquid is constant. Derive the differential equations in terms of the liquid heights h_1 and h_2 . Write the equations in second-order matrix form.

6. [Figure 7.18](#) shows a liquid-level system in which two tanks have cross-sectional areas A_1 and A_2 , respectively. The volume flow rate into tank 1 is q_i . A pump is connected to the bottom of tank 1, and the pressure of the fluid increases by Δp when crossing the pump. Tank 2 is located higher than tank 1, and the vertical distance between the two tanks is H . The liquid is pumped from tank 1 to tank 2 through a valve of linear resistance R_1 and leaves tank 2 through a valve of linear resistance R_2 . The density ρ of the liquid is constant. Derive the differential equations in terms of the liquid heights h_1 and h_2 . Write the equations in second-order matrix form.
7. Consider the single-tank liquid-level system shown in [Figure 7.19](#), where the volume flow rate into the tank through a pipe is q_i . The liquid leaves the tank through an orifice of area A_o . Denote C_d as the discharge coefficient, which is the ratio of the actual mass flow rate to the theoretical one, and lies in the range of $0 < C_d < 1$ because of friction effects. Derive the differential equation relating the liquid

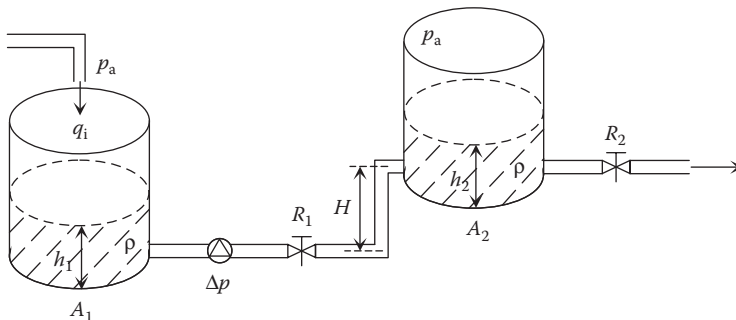


FIGURE 7.18
Problem 6.

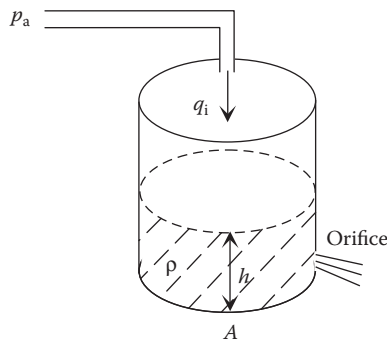


FIGURE 7.19
Problem 7.

height h and the volume flow rate q_i at the inlet. The tank's cross-sectional area is constant. The density ρ of the liquid is constant.

8. Figure 7.20 shows a hydraulic system of two interconnected tanks that have the same cross-sectional area of A . A pump is connected to tank 1. Assume that the relationship between the voltage applied to the pump and the mass flow rate into tank 1 is linear; that is, $q_{mi} = k_p v_a$, where k_p is called the pump constant and can be obtained by experimental measurements. Tank 1 is connected to tank 2, which is connected to a reservoir. The liquid leaves each tank through an outlet of area A_o at the bottom. Derive the differential equations in terms of the liquid heights h_1 and h_2 .

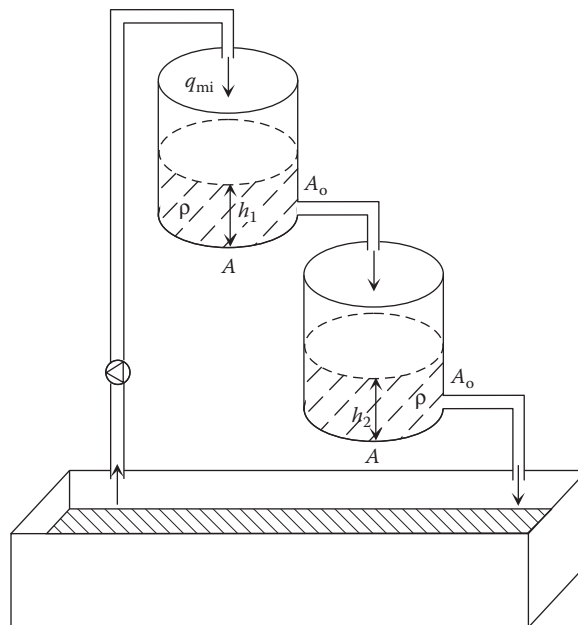


FIGURE 7.20
Problem 8.

7.3 Thermal Systems

Thermal systems are those that involve the transfer of heat from one object to another. When an object is at a different temperature than its surroundings or another object, heat will transfer from the higher-temperature object to the lower-temperature one, obeying the law of conservation of energy. Examples of thermal systems include heaters, air conditioners, refrigerators, and so on. As with fluid systems (either pneumatic or hydraulic), for which fluid capacitance, fluid resistance, and the conservation of mass are the basis for system modeling, we will introduce thermal capacitance, thermal resistance, and the conservation of energy, which together form the basis of modeling for thermal systems.

7.3.1 First Law of Thermodynamics

The first law of thermodynamics is a version of the law of conservation of energy, adapted for thermodynamic systems. For a system with well-defined boundaries, the law of conservation of energy states

$$\Delta E = Q - W \quad (7.37)$$

where:

- ΔE is the change in energy of the system
- Q is the heat flow into or out of the system
- W is the work done by or on the system

In Equation 7.37, Q is positive if heat is supplied to the system and negative if heat is given off by the system. W is positive if work is done by the system and negative if work is done on the system. Based on this sign convention, we have

$$\Delta E = (Q_{\text{in}} - Q_{\text{out}}) - (W_{\text{out}} - W_{\text{in}}) \quad (7.38)$$

where Q_{in} , Q_{out} , W_{in} , and W_{out} are all positive quantities.

In actuality, the net amount of energy added to the system is equal to the net increase in the energy stored internally in the system and any change in the mechanical energy of the system's center of mass,

$$\Delta E = \Delta U + \Delta ME_C \quad (7.39)$$

U is the internal energy (or internal thermal energy), which is the energy stored at the molecular level. It includes the kinetic energy due to the motion of molecules and the potential energy that holds the atoms together. ME_C stands for the mechanical energy, which includes the kinetic energy and the potential energy associated with the system's center of mass. For systems with negligible change in mechanical energy,

$$\Delta U = Q - W = (Q_{\text{in}} - Q_{\text{out}}) - (W_{\text{out}} - W_{\text{in}}) \quad (7.40)$$

which is the mathematical expression of the first law of thermodynamics. It basically states that the change in internal energy is equal to the amount of energy gained by heating minus the amount lost by doing work on the environment.

Heat Q is the energy transfer at the molecular level. Work W is the energy transfer that is capable of producing macroscopic mechanical motion of the system's center of mass. For thermal systems with pure heat transfer and no work involved, that is, $W_{\text{in}} = W_{\text{out}} = 0$, the law of conservation of energy presented in Equation 7.40 can be rewritten as

$$\Delta U = Q = Q_{\text{in}} - Q_{\text{out}} \quad (7.41)$$

or

$$\frac{dU}{dt} = q_{\text{hi}} - q_{\text{ho}} \quad (7.42)$$

where $q_{\text{h}} = dQ/dt$ is the heat flow rate and has the units of J/s, which is a watt or ft·lb/s.

7.3.2 Thermal Capacitance

For an object, the thermal capacitance C is defined as the ratio of the change in heat flow to the change in the object's temperature,

$$C = \frac{dQ}{dT} \quad (7.43)$$

where C has the units of J/K, J/°C, or ft·lb/°F. The thermal capacitance is a measure of the heat required to increase the temperature of an object by a certain temperature interval.

Strictly speaking, the value of the thermal capacitance of a substance depends on thermodynamic processes. For a constant-volume process, no work is involved and all the heat goes into the internal energy of the substance,

$$Q = \Delta U = mc_v \Delta T \quad (7.44)$$

where:

m is the mass of the substance

c_v is the constant-volume specific heat capacity of the substance in the units of J·K/kg, J·°C/kg, or ft·lb·°F/slug

ΔT is the change in temperature of the substance

For a constant-pressure process,

$$Q = \Delta H = mc_p \Delta T \quad (7.45)$$

where:

H is the enthalpy

c_p is the constant-pressure specific heat capacity

Combining Equation 7.43 with Equation 7.44 or 7.45 gives

$$C = mc_v \quad (7.46)$$

or

$$C = mc_p \quad (7.47)$$

For incompressible liquids and solids, because the volume cannot expand, the heat flow in a constant-pressure process is equal to the internal energy: $Q = \Delta U$ and $c_p = c_v$. Assuming that the density and the volume of the mass are ρ and V , respectively, we have

$$C = mc = \rho Vc \quad (7.48)$$

where c is the specific heat capacity and $c = c_p = c_v$. The subscripts p and v will be dropped in the rest of the chapter for simplicity.

Note that the value of the specific heat capacity c depends on the substance of the object, whereas the thermal capacitance C is an extensive property, because its value is proportional to the mass of the object. For instance, the specific heat capacity of water at room temperature (25°C) is a constant value of 4186 J/(kg·°C). However, the thermal capacitance for a bathtub of water is greater than that for a cup of water.

7.3.3 Thermal Resistance

There are three mechanisms by which heat is transported: conduction, convection, and radiation. Conduction is the transfer of heat between substances that are in direct contact with each other. Convection is the transfer of heat due to a flowing fluid, which can be a gas or a liquid. Radiation is the transfer of heat through empty space. Here, we only consider conduction and convection.

The thermal resistance R for heat transfer is defined as the ratio of the change in temperature difference to the change in heat flow rate,

$$R = \frac{dT}{dq_h} \quad (7.49)$$

The thermal resistance R has units of Ks/J, °Cs/J, or °F·s/(ft·lb).

For simple one-dimensional conduction, as shown in [Figure 7.21](#), Fourier's law, also known as the law of heat conduction, gives

$$q_h = kA \frac{\Delta T}{L} = kA \frac{T_1 - T_2}{L} \quad (7.50)$$

where:

L is the length of the body in the direction of heat flow

A is the cross-sectional area normal to the direction of heat flow

ΔT is the temperature difference along its length

k is the thermal conductivity of the material in W/(m·K) or W/(m·°C)

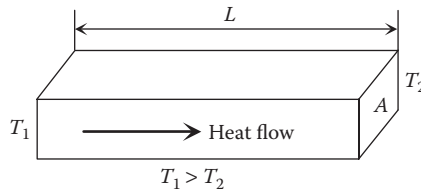


FIGURE 7.21

One-dimensional conduction: heat flow from higher to lower temperature.

Note that the heat flow is in the direction of decreasing temperature. Fourier's equation is often used only for solids; however, it is valid for both solids and fluids. Combining Equations 7.49 and 7.50 gives the thermal resistance for conduction

$$R = \frac{L}{kA} \quad (7.51)$$

For convective heat transfer, Newton's law of cooling states that the rate of heat flow of a body is proportional to the difference in temperatures between the body and its surroundings or environment. The mathematical expression is

$$q_h = hA\Delta T = hA(T_s - T_{\text{env}}) \quad (7.52)$$

where:

A is the surface area from which the heat is being transferred

T_s is the temperature of the body's surface

T_{env} is the temperature of the environment

h is the heat transfer coefficient in $W/(m^2 \cdot K)$ or $W/(m^2 \cdot ^\circ C)$

Combining Equations 7.49 and 7.52 gives the thermal resistance for convection

$$R = \frac{1}{hA} \quad (7.53)$$

It is very useful to utilize the concept of thermal resistance and represent heat transfer by thermal circuits. The heat flow rate q_h is analogous to the current, the temperature difference ΔT is analogous to the voltage, and the thermal resistance is analogous to the electric resistance.

Figure 7.22 shows the heat transfer across a composite slab, which can be represented using a thermal resistance network with series interconnection. The heat flow rate remains the same through each component. Assume that the thermal resistances are R_1 and R_2 . As a result, the temperature differences across the resistances R_1 and R_2 are $T_1 - T_2 = R_1 q_h$ and $T_2 - T_3 = R_2 q_h$, respectively. The total temperature difference across the composite slab is $T_1 - T_3 = (R_1 + R_2) q_h$. Thus, the equivalent thermal resistance for a series interconnection is

$$R_{\text{eq}} = R_1 + R_2 \quad (7.54)$$

Figure 7.23 shows heat transfer across a wall with different materials and how it can be represented by using a thermal resistance network with parallel interconnection. Note that the temperature difference across each material must be the same. Consequently, the heat flow rates through the resistances R_1 and R_2 are $q_{h1} = (T_1 - T_2)/R_1$ and $q_{h2} = (T_1 - T_2)/R_2$, respectively. The total heat flow rate through the wall is $q_{h1} + q_{h2} = (T_1 - T_2)(1/R_1 + 1/R_2)$. Thus, the equivalent thermal resistance for a parallel interconnection is

$$\frac{1}{R_{\text{eq}}} = \frac{1}{R_1} + \frac{1}{R_2} \quad (7.55)$$

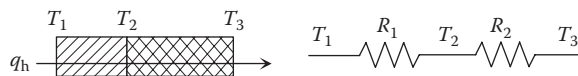
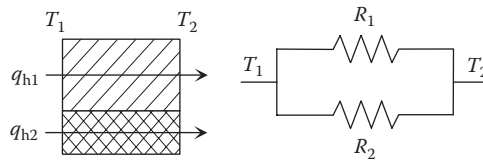


FIGURE 7.22

Heat transfer across series thermal resistance.

**FIGURE 7.23**

Heat transfer across parallel thermal resistance.

If there are different heat transfer modes in a system, a thermal circuit with thermal resistances representing the different modes of heat transfer can be used to analyze the system.

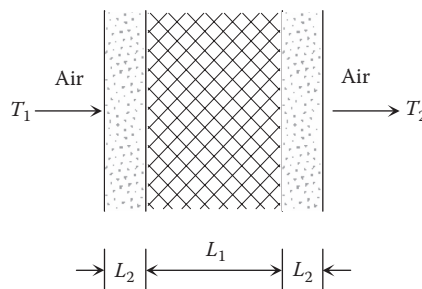
Example 7.6: Thermal Resistance

Consider heat transfer through an insulated wall, as shown in Figure 7.24. The wall is made of a layer of brick with thermal conductivity k_1 and two layers of foam with thermal conductivity k_2 for insulation. The left surface of the wall is at temperature T_1 and exposed to air, with heat transfer coefficient h_1 . The right surface of the wall is at temperature T_2 and exposed to air, with heat transfer coefficient h_2 . Assume that $k_1 = 0.5 \text{ W}/(\text{m}^\circ\text{C})$, $k_2 = 0.17 \text{ W}/(\text{m}^\circ\text{C})$, $h_1 = h_2 = 10 \text{ W}/(\text{m}^2\text{C})$, $T_1 = 38^\circ\text{C}$, and $T_2 = 20^\circ\text{C}$. The thickness of the brick layer is 0.1 m, the thickness of each foam layer is 0.03 m, and the cross-sectional area of the wall is 16 m^2 .

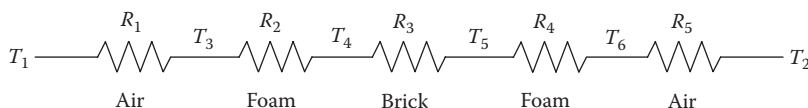
- Determine the heat flow rate through the wall.
- Determine the temperature distribution through the wall.

Solution

- The heat transfer through the insulated wall can be represented using a thermal circuit with five thermal resistances connected in series, as shown in Figure 7.25. Two modes of heat transfer, conduction and convection, are involved. The corresponding thermal resistances are

**FIGURE 7.24**

Heat transfer through an insulated wall.

**FIGURE 7.25**

The equivalent thermal circuit for the heat transfer system in Figure 7.24.

$$R_1 = \frac{1}{h_1 A} = \frac{1}{10 \times 16} = 6.25 \times 10^{-3} \text{ } ^\circ\text{C} \cdot \text{s/J}$$

$$R_2 = R_4 = \frac{L_2}{k_2 A} = \frac{0.03}{0.17 \times 16} = 1.10 \times 10^{-2} \text{ } ^\circ\text{C} \cdot \text{s/J}$$

$$R_3 = \frac{L_1}{k_1 A} = \frac{0.1}{0.5 \times 16} = 1.25 \times 10^{-2} \text{ } ^\circ\text{C} \cdot \text{s/J}$$

$$R_5 = \frac{1}{h_2 A} = \frac{1}{10 \times 16} = 6.25 \times 10^{-3} \text{ } ^\circ\text{C} \cdot \text{s/J}$$

The total thermal resistance is

$$R_{\text{eq}} = \sum_{i=1}^5 R_i = 4.70 \times 10^{-2} \text{ } ^\circ\text{C} \cdot \text{s/J}$$

Thus, the heat flow rate through the insulated wall is

$$q_h = \frac{\Delta T}{R_{\text{eq}}} = \frac{38 - 20}{4.70 \times 10^{-2}} = 382.98 \text{ W}$$

- b. Note that the heat flow rate stays the same through the insulated wall. Thus, from left to right, the heat flow rate through each layer is

$$\text{Air: } q_h = R_1(T_1 - T_3)$$

$$\text{Foam: } q_h = R_2(T_3 - T_4)$$

$$\text{Brick: } q_h = R_3(T_4 - T_5)$$

$$\text{Foam: } q_h = R_4(T_5 - T_6)$$

$$\text{Air: } q_h = R_5(T_6 - T_2)$$

With the given temperatures T_1 and T_2 , we have $T_3 = 35.61^\circ\text{C}$, $T_4 = 31.40^\circ\text{C}$, $T_5 = 26.61^\circ\text{C}$, and $T_6 = 22.40^\circ\text{C}$. [Figure 7.26](#) shows the temperature distribution through the wall. Note that the temperatures shown in [Figure 7.26](#) are the values when the heat transfer process reaches a steady state.

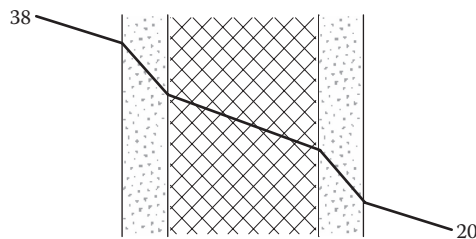


FIGURE 7.26

Temperature distribution through the insulated wall.

7.3.4 Modeling of Heat Transfer Systems

The mathematical model of a thermal system is often complicated because of the complex temperature distribution throughout the system. To simplify analysis, in this section, we discuss how to obtain a lumped-parameter model, which may approximate the gross system dynamics. The validity of this lumped-parameter assumption can be checked using the so-called Biot number.

Consider a solid body submerged in hot fluid, as shown in [Figure 7.27](#), in which T_f is the fluid temperature, T_w is the temperature at the wall of the object, and T is the temperature of an arbitrary point inside the object. There are two modes of heat transfer involved, conduction within the body and convection between the fluid and the body. The heat flow rates for the two different modes can be approximated to have the same magnitude,

$$\frac{kA}{L}(T_w - T) \approx hA(T_f - T_w) \quad (7.56)$$

where:

h is the heat transfer coefficient

k is the thermal conductivity for the material of the object

L is the relevant length between the wall and the point

The ratio of the temperature differences caused by the different modes of heat transfer is

$$\frac{T_w - T}{T_f - T_w} \approx \frac{hL}{k} \quad (7.57)$$

Note that the temperature difference within the object can be negligible if the object is thin or small enough. The criterion for determining a solid being thin or small is based on the Biot number, which is defined as

$$Bi = \frac{hL_c}{k} \quad (7.58)$$

where L_c is the characteristic length of the solid object. The characteristic length is defined as the volume of the body divided by the surface area of the body,

$$L_c = \frac{V_{\text{body}}}{A_{\text{surface}}} \quad (7.59)$$

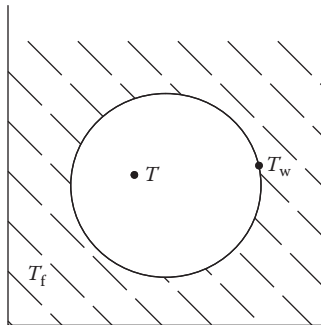


FIGURE 7.27

A solid object submerged in a fluid.

For a body whose Biot number is much less than one, typically, $Bi < 0.1$, the interior of the body may be assumed to have the same temperature.

Example 7.7: Temperature Dynamics of a Heated Object

Consider a steel sphere with a radius of $r = 0.01$ m submerged in a hot water bath, with a heat transfer coefficient of $h = 350$ W/(m²·°C). For steel, the density is $\rho = 7850$ kg/m³, the specific heat capacity is $c = 440$ J/(kg·°C), and the thermal conductivity is $k = 43$ W/(m·°C). The temperature of the water T_f is 100°C, and the initial temperature of the sphere T_0 is 25°C.

- Determine whether the sphere's temperature can be considered uniform.
- Derive the differential equation relating the sphere's temperature $T(t)$ and the water's temperature T_f .
- Using the differential equation obtained in Part (b), construct a Simulink block diagram to find the sphere's temperature $T(t)$.

Solution

- The characteristic length of the sphere is

$$L_c = \frac{V_{\text{body}}}{A_{\text{surface}}} = \frac{\frac{4}{3}\pi r^3}{4\pi r^2} = \frac{1}{3}r = \frac{0.01}{3}$$

The Biot number of the steel sphere is

$$Bi = \frac{hL_c}{k} = \frac{350(0.01)}{43(3)} = 2.71 \times 10^{-2} < 0.1$$

Thus, the sphere can be treated as a lump-temperature system, and its temperature can be considered uniform within the body.

- The dynamic model of the sphere's temperature can be derived using the law of conservation of energy,

$$\frac{dU}{dt} = q_{\text{hi}} - q_{\text{ho}}$$

Note that $U = mcT = \rho VcT$, and we have

$$\frac{dU}{dt} = \frac{d}{dt}(\rho VcT) = \rho Vc \frac{dT}{dt}$$

The heat flow rate into the body is

$$q_{\text{hi}} = \frac{T_f - T}{R}$$

and the heat flow rate out of the body is $q_{\text{ho}} = 0$. Thus, the differential equation of the system is

$$\rho Vc (dT/dt) = (T_f - T)/R$$

Introducing the expression for the thermal capacitance $\rho Vc = C$, gives

$$RC \frac{dT}{dt} + T = T_f$$

The thermal capacitance of the sphere is

$$C = \rho V c = 7850 \left(\frac{4}{3} \right) (\pi) (0.01)^3 (440) = 14.47 \text{ J/}^\circ\text{C}$$

and the thermal resistance due to convection is

$$R = \frac{1}{hA} = \frac{1}{350(4)(\pi)(0.01)^2} = 2.27^\circ\text{C s/J}$$

Thus, the dynamic model of the sphere's temperature is

$$32.85 \frac{dT}{dt} + T = T_i$$

- c.  Given $T_i = 100^\circ\text{C}$, solving for the highest derivative of the output T gives

$$\frac{dT}{dt} = \frac{1}{32.85} (100 - T)$$

which can be represented using the block diagram shown in [Figure 7.28](#). Double-click on the Integrator block and define the initial temperature of the sphere as 25°C . Run the simulation. The results can be plotted as shown in [Figure 7.29](#).

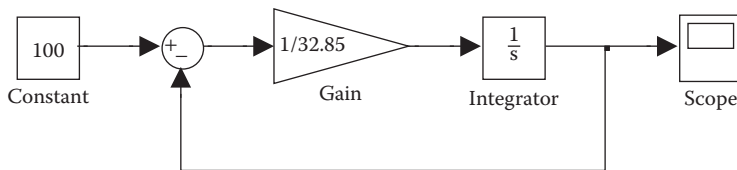


FIGURE 7.28

A Simulink block diagram representing the thermal system in Example 7.7.

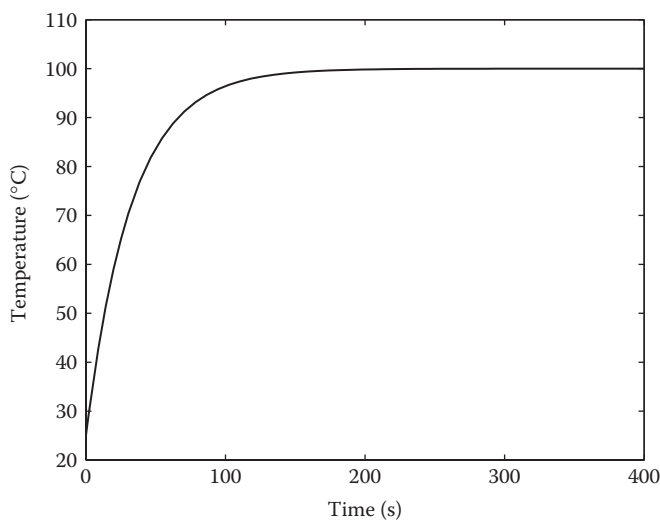


FIGURE 7.29

Temperature output $T(t)$ of the thermal system in Example 7.7.

Example 7.7 shows that the temperature dynamics of a thermal system can be expressed in terms of thermal capacitance and thermal resistance. For multiple thermal capacitances, we apply the law of conservation of energy to each of them.

Example 7.8: Temperature Dynamics of a House with Double-Layer Walls

As shown in [Figure 7.30](#), the wall of a room consists of two layers, for which the thermal capacitances are C_1 and C_2 . Assume that the temperatures in both layers are uniform and they are T_1 and T_2 , respectively. The temperatures inside and outside the room are T_i and T_o , respectively. Both layers exchange heat by convection with air and the thermal resistances are R_1 and R_3 , respectively. The thermal resistance of the interface between the layers is R_2 .

- Derive the differential equations for this system.
- Using the differential equations obtained in Part (a), determine the state-space form of the system. Assume the temperatures T_1 and T_2 are the outputs.

Solution

- Assume that the temperature inside the room is lower than the one outside. Applying the law of conservation of energy to the outer layer, we have

$$\frac{dU}{dt} = q_{hi} - q_{ho}$$

where

$$\frac{dU}{dt} = C_1 \frac{dT_1}{dt}$$

$$q_{hi} = \frac{T_o - T_1}{R_1}$$

$$q_{ho} = \frac{T_1 - T_2}{R_2}$$

Substituting these expressions gives

$$C_1 \frac{dT_1}{dt} + \left(\frac{1}{R_1} + \frac{1}{R_2} \right) T_1 - \frac{1}{R_2} T_2 = \frac{1}{R_1} T_o$$

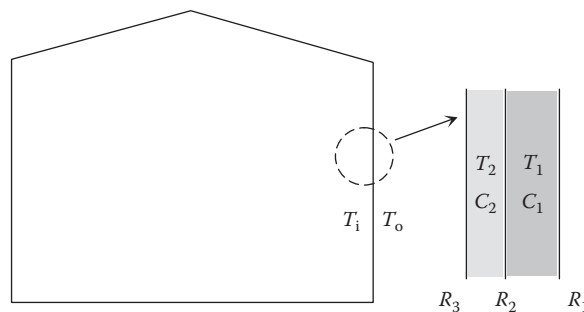


FIGURE 7.30
A room with double-layer walls.

Applying the law of conservation of energy to the inner layer, we have

$$\frac{dU}{dt} = q_{hi} - q_{ho}$$

where

$$\frac{dU}{dt} = C_2 \frac{dT_2}{dt}$$

$$q_{hi} = \frac{T_1 - T_2}{R_2}$$

$$q_{ho} = \frac{T_2 - T_i}{R_3}$$

Substituting these expressions gives

$$C_2 \frac{dT_2}{dt} - \frac{1}{R_2} T_1 + \left(\frac{1}{R_2} + \frac{1}{R_3} \right) T_2 = \frac{1}{R_3} T_i$$

The system of differential equations can be written in matrix form as

$$\begin{bmatrix} C_1 & 0 \\ 0 & C_2 \end{bmatrix} \begin{Bmatrix} \frac{dT_1}{dt} \\ \frac{dT_2}{dt} \end{Bmatrix} + \begin{bmatrix} \left(\frac{1}{R_1} + \frac{1}{R_2} \right) & -\frac{1}{R_2} \\ -\frac{1}{R_2} & \left(\frac{1}{R_2} + \frac{1}{R_3} \right) \end{bmatrix} \begin{Bmatrix} T_1 \\ T_2 \end{Bmatrix} = \begin{Bmatrix} \frac{1}{R_1} T_o \\ \frac{1}{R_3} T_i \end{Bmatrix}$$

- b. To represent a thermal system in the state-space form, the temperature of each thermal capacitance is often chosen as a state variable. As specified, the state, the input, and the output are

$$\mathbf{x} = \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix} = \begin{Bmatrix} T_1 \\ T_2 \end{Bmatrix}, \quad \mathbf{u} = \begin{Bmatrix} T_i \\ T_o \end{Bmatrix}, \quad \mathbf{y} = \begin{Bmatrix} T_1 \\ T_2 \end{Bmatrix}$$

The state-variable equations are

$$\dot{x}_1 = \frac{dT_1}{dt} = -\left(\frac{1}{R_1 C_1} + \frac{1}{R_2 C_1} \right) T_1 + \frac{1}{R_2 C_1} T_2 + \frac{1}{R_1 C_1} T_o$$

$$\dot{x}_2 = \frac{dT_2}{dt} = \frac{1}{R_2 C_2} T_1 - \left(\frac{1}{R_2 C_2} + \frac{1}{R_3 C_2} \right) T_2 + \frac{1}{R_3 C_2} T_i$$

or in matrix form

$$\begin{Bmatrix} \dot{x}_1 \\ \dot{x}_2 \end{Bmatrix} = \begin{bmatrix} -\left(\frac{1}{R_1 C_1} + \frac{1}{R_2 C_1} \right) & \frac{1}{R_2 C_1} \\ \frac{1}{R_2 C_2} & -\left(\frac{1}{R_2 C_2} + \frac{1}{R_3 C_2} \right) \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix} + \begin{bmatrix} 0 & \frac{1}{R_1 C_1} \\ \frac{1}{R_3 C_2} & 0 \end{bmatrix} \begin{Bmatrix} u_1 \\ u_2 \end{Bmatrix}$$

The output equation is

$$\begin{Bmatrix} y_1 \\ y_2 \end{Bmatrix} = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix} + \begin{bmatrix} 0 & 0 \\ 0 & 0 \end{bmatrix} \begin{Bmatrix} u_1 \\ u_2 \end{Bmatrix}$$

PROBLEM SET 7.3

1. Consider heat transfer through an insulated frame wall of a house. The thermal conductivity of the wall is $0.055 \text{ W}/(\text{m}\cdot^\circ\text{C})$. The wall is 0.15 m thick and has an area of 15 m^2 . The inside air temperature is 20°C and the heat transfer coefficient for convection between the wall and the inside air is $2.6 \text{ W}/(\text{m}^2\cdot^\circ\text{C})$. On the outside of the wall, the heat transfer coefficient for convection between the wall and the outside air is $10.4 \text{ W}/(\text{m}^2\cdot^\circ\text{C})$ and the outside air temperature is -10°C . Determine the heat flow rate through the wall.
2. Consider heat transfer through a double-pane window as shown in [Figure 7.31a](#). Two layers of glass with thermal conductivity k_1 are separated by a layer of stagnant air with thermal conductivity k_2 . The inner surface of the window is at temperature T_1 and exposed to room air with heat transfer coefficient h_1 . The outer surface of the wall is at temperature T_2 and exposed to air with heat transfer coefficient h_2 . Assume that $k_1 = 0.95 \text{ W}/(\text{m}\cdot^\circ\text{C})$, $k_2 = 0.0285 \text{ W}/(\text{m}\cdot^\circ\text{C})$, $h_1 = h_2 = 10 \text{ W}/(\text{m}^2\cdot^\circ\text{C})$, $T_1 = 20^\circ\text{C}$, and $T_2 = 35^\circ\text{C}$. The thickness of each glass layer is 4 mm , the thickness of the air layer is 8 mm , and the cross-sectional area of the window is 1.5 m^2 .
 - a. Determine the heat flow rate through the double-pane window.
 - b. Determine the temperature distribution through the double-pane window.
 - c. Repeat Parts (a) and (b) for the single-pane glass window shown in [Figure 7.31b](#).
3. The junction of a thermocouple can be approximated as a sphere with a diameter of 1 mm . As shown in [Figure 7.32](#), the thermocouple is used to measure the temperature of a gas stream. For the junction, the density is $\rho = 8500 \text{ kg}/\text{m}^3$, the specific heat capacity is $c = 320 \text{ J}/(\text{kg}\cdot^\circ\text{C})$, and the thermal conductivity is $k = 40 \text{ W}/(\text{m}\cdot^\circ\text{C})$. The temperature of the gas T_f is 120°C and the initial temperature of the

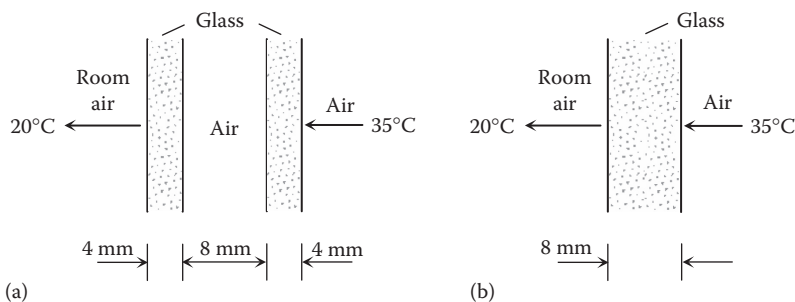


FIGURE 7.31
Problem 2.

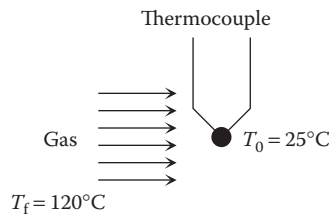


FIGURE 7.32
Problem 3.

sphere T_0 is 25°C . The heat transfer coefficient between the gas and the junction is $h = 70 \text{ W}/(\text{m}^2\cdot^\circ\text{C})$.

- Determine if the junction's temperature can be considered uniform.
 - Derive the differential equation relating the junction's temperature $T(t)$ and the gas's temperature T_f .
 - Using the differential equation obtained in Part (b), construct a Simulink block diagram to find out how long it will take the thermocouple to read 99% of the initial temperature difference.
4. Figure 7.33 shows a thin-walled glass of milk, which is taken out of the refrigerator at a uniform temperature of 3°C and is placed in a large pan filled with hot water at 60°C . Assume that the assumption of the lumped system analysis is applicable because the milk is stirred constantly, so that its temperature is uniform at all times. The glass container is cylindrical in shape, with a radius of 3 cm and a height of 6 cm. The estimated parameters of the milk are density $\rho = 1035 \text{ kg}/\text{m}^3$, specific heat capacity $c = 3980 \text{ J}/(\text{kg}\cdot^\circ\text{C})$, and thermal conductivity $k = 0.56 \text{ W}/(\text{m}\cdot^\circ\text{C})$. The heat transfer coefficient between the water and the glass is $h = 250 \text{ W}/(\text{m}^2\cdot^\circ\text{C})$.
- Derive the differential equation relating the milk temperature $T(t)$ and the water temperature.
 - Using the differential equation obtained in Part (a), construct a Simulink block diagram. How long will it take for the milk to warm up from 3°C to 58°C ?

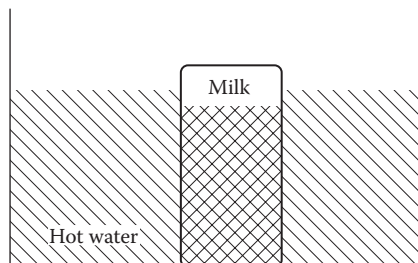


FIGURE 7.33
Problem 4.

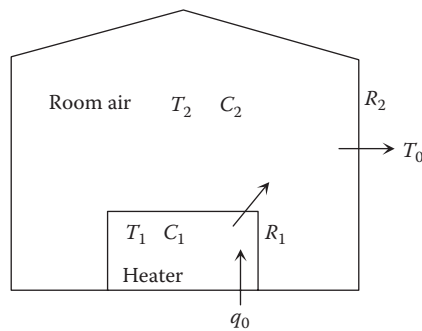


FIGURE 7.34
Problem 5.

5. The room shown in Figure 7.34 has a heater with heat flow rate input of q_0 . The thermal capacitances of the heater and the room air are C_1 and C_2 , respectively. The thermal resistances of the heater–air interface and the room wall–ambient air interface are R_1 and R_2 , respectively. The temperatures of the heater and the room air are T_1 and T_2 , respectively. The temperature outside the room is T_0 , which is assumed to be constant.
 - a. Derive the differential equations relating the temperatures T_1 , T_2 , the input q_0 , and the outside temperature T_0 .
 - b. Using the differential equations obtained in Part (a), determine the state-space form of the system. Assume the temperatures T_1 and T_2 as the outputs.
6. For the three-room house shown in Figure 7.35, all rooms are perfect squares and have the same dimensions. An air conditioner produces an equal amount of heat flow q_{ho} out of each room. The temperature outside the house is T_0 . Assume that there is no heat flow through the floors or ceilings. The thermal resistances through the inner walls and the outer walls are R_i and R_o , respectively. The thermal capacitance of each room is C . Derive the differential equations for this system.

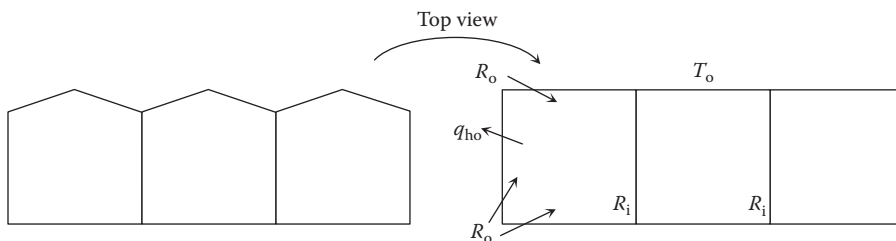


FIGURE 7.35
Problem 6.

7.4 System Modeling with Simulink and Simscape

As in the modeling of mechanical and electrical systems, the dynamics of fluid and thermal systems can be represented by ordinary differential equations, transfer functions, or state-space form. Therefore, the Simulink modeling techniques discussed in [Sections 5.6](#) and [6.6](#) can also be applied to fluid and thermal systems.

Example 7.9: Modeling of a Pneumatic System with Simulink

Consider the pneumatic system in Example 7.2. Construct a Simulink block diagram to find the pressure inside the container, $p(t)$, which is assumed to be 0 Pa initially. The pressure at the inlet is assumed to be 101.325 kPa.

Solution

The dynamics equation obtained in Example 7.2 is

$$\frac{RV}{R_{\text{air}}T} \frac{dp}{dt} + p = p_i$$

where $RV/(R_{\text{air}}T) = 1.19 \times 10^{-2}$ s. Solving for the highest derivative of the output p gives

$$\dot{p} = 84.03(101,325 - p)$$

which can be represented by the block diagram shown in [Figure 7.36](#). One Integrator block is used to form the container pressure p , which is fed back to form the variation rate $\dot{p}$. Note that the system input is the inlet pressure p_i , which is constant and is represented by a Constant block. Run the simulation. Double-click the Scope block; the resulting output of the pneumatic system $p(t)$ is shown in [Figure 7.37](#).

Example 7.10: Modeling of a Two-Tank Liquid-Level System with Simulink

Consider the two-tank liquid-level system in Example 7.5. Construct a Simulink block diagram to find the liquid levels $h_1(t)$ and $h_2(t)$. Assume that $\rho = 1000 \text{ kg/m}^3$, $g = 9.81 \text{ m/s}^2$, $A_1 = 2 \text{ m}^2$, $A_2 = 3 \text{ m}^2$, $R_1 = R_2 = R_3 = 400 \text{ N}\cdot\text{s}/(\text{kg}\cdot\text{m}^2)$, and initial liquid heights $h_1(0) = 1 \text{ m}$ and $h_2(0) = 0 \text{ m}$. The pump pressure Δp is a step function with a magnitude of 0 before $t = 0 \text{ s}$ and a magnitude of 130 kPa after $t = 0 \text{ s}$.

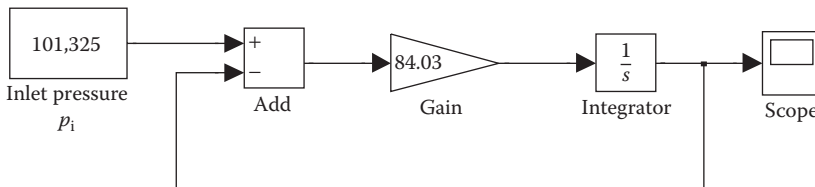


FIGURE 7.36
Simulink block diagram representing the pneumatic system in Example 7.2.

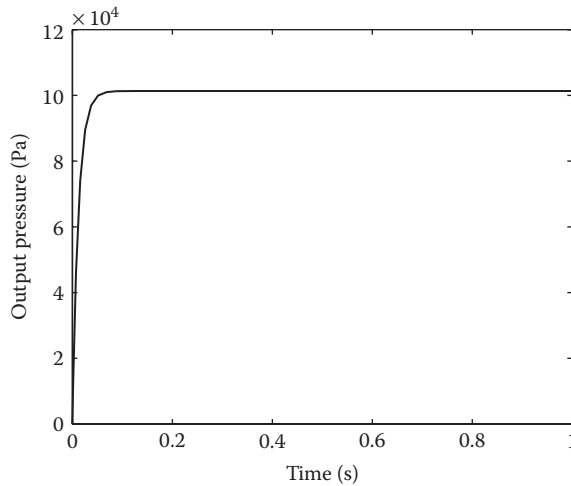


FIGURE 7.37
Pressure output $p(t)$ for constant inlet pressure $p_i = 101.325$ kPa.

Solution

The Simulink block diagram can be constructed based on either the differential equations obtained in Part (a) or the state-space form obtained in Part (b) in Example 7.5. Substituting the parameter values into the differential equations gives

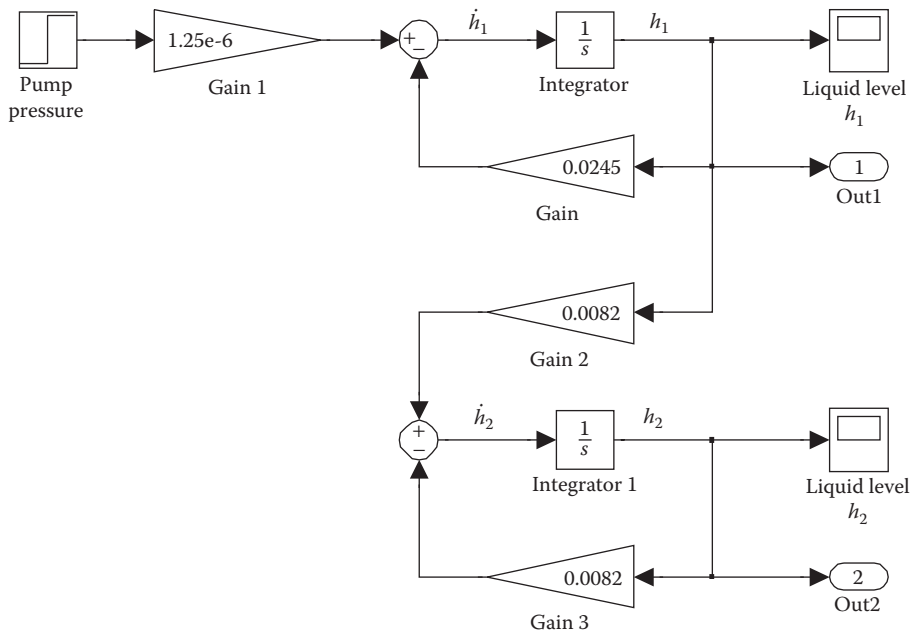
$$\begin{aligned}\frac{dh_1}{dt} &= -0.0245h_1 + 1.25 \times 10^{-6} \Delta p \\ \frac{dh_2}{dt} &= 0.0082h_1 - 0.0082h_2\end{aligned}$$

Figure 7.38 shows the resulting Simulink block diagram, where two Integrator blocks are used to form h_1 and h_2 . By double-clicking each Integrator block, we can enter the initial liquid level for each tank.

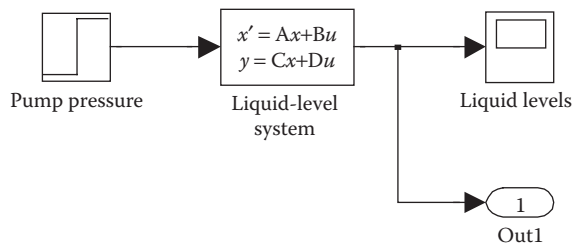
Substituting the parameter values into the state-space equations gives

$$\begin{aligned}\begin{Bmatrix} \dot{x}_1 \\ \dot{x}_2 \end{Bmatrix} &= \begin{bmatrix} -0.0245 & 0 \\ 0.0082 & -0.0082 \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix} + \begin{bmatrix} 1.25 \times 10^{-6} \\ 0 \end{bmatrix} u \\ \begin{Bmatrix} y_1 \\ y_2 \end{Bmatrix} &= \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix} + \begin{bmatrix} 0 \\ 0 \end{bmatrix} u\end{aligned}$$

The Simulink block diagram based on the state-space form is shown in Figure 7.39, where a State-Space block is used to represent the liquid-level system. By double-clicking the State-Space block with the name `Liquid-level system`, we can define the

**FIGURE 7.38**

Simulink block diagram constructed based on differential equations.

**FIGURE 7.39**

Simulink block diagram constructed based on state-space equations.

matrices **A**, **B**, **C**, and **D**. The initial liquid level is a vector, which is $[1; 0]$. By running either of the two simulations, we can obtain the same results as plotted in [Figure 7.40](#).

To model a fluid or thermal system with Simscape, the Simscape\Foundation Library can be used, which contains basic pneumatic, hydraulic, and thermal blocks. However, connecting pneumatic or hydraulic components requires a good knowledge of the physical domains and the equations involved. Because this is beyond the scope of this text, this section focuses on Simscape modeling of thermal systems only.

Example 7.11: Temperature Dynamics of a Heated Object

Consider the heat transfer system in Example 7.7, in which a steel sphere is submerged in hot water and the temperature of the sphere is assumed to be uniform. Build a Simscape model of the physical system, and find the sphere's temperature output $T(t)$. Compare the result with that obtained in Example 7.7.

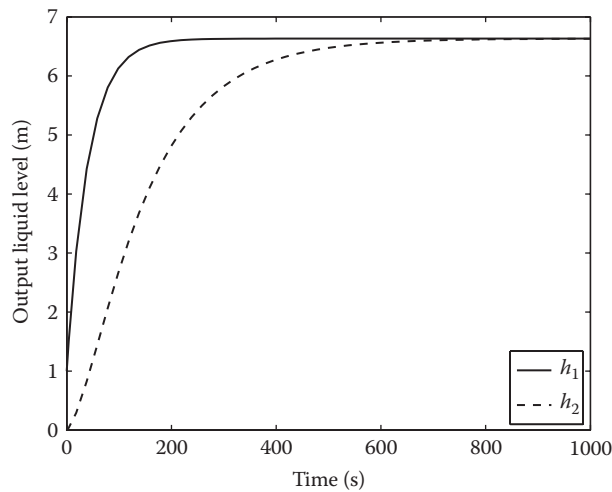


FIGURE 7.40
Liquid level outputs $h_1(t)$ and $h_2(t)$.

Solution

The Simscape block diagram corresponding to the physical system is shown in Figure 7.41.

The temperature of hot water is 100°C or 373 K . The Ideal Temperature Source block in the library of Simscape/Foundation Library/Thermal/Thermal Sources is used to represent the temperature input. A Simulink-PS Converter block converts the constant value of 373 K into a physical signal. Double-click the block, and define the Input signal unit as K .

Because the temperature of the sphere is assumed to be uniform, only the convective heat transfer between the water and the sphere is considered in modeling. The corresponding block can be found in the library of Simscape/Foundation Library/Thermal/Thermal Elements. Double-click the block, and define Area as 0.0013 m^2 and Heat transfer coefficient as $350\text{ W}/(\text{m}^2\cdot\text{K})$. The block Thermal Mass in the same library is used to represent the steel sphere. The associated parameters are Mass, Specific heat, and Initial temperature, and their values are 0.0329 kg ,

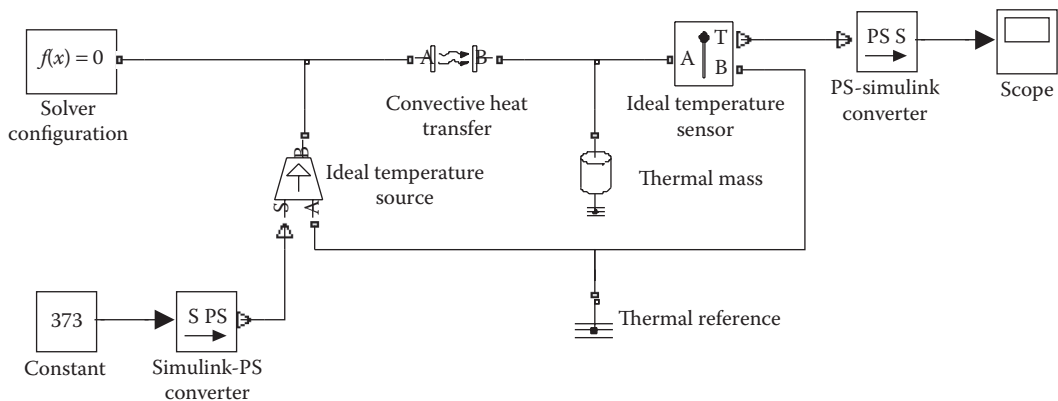


FIGURE 7.41
Simscape block diagram of the thermal system in Example 7.7.

440 J/(kg·K), and 25°C or 298 K. All parameter values can be determined using the information given in Example 7.7.

To measure the sphere's temperature, drag the Ideal Temperature Sensor block in the library of Simscape/Foundation Library/Thermal/Thermal Sensors. The PS-Simulink Converter block converts the physical signal to a Simulink signal. Double-click the block, and define the Output signal unit as K. Since the simulation result in Example 7.7 is given in the unit of °C, we can also define the unit as C and check the box of Apply affine conversion.

Run the simulation, and the same curve as shown in Figure 7.29 can be generated, which is the resulting temperature output $T(t)$ of the heated sphere.

PROBLEM SET 7.4

1. Dry air at a constant temperature of 25°C passes through a valve out of a rigid cubic container of 1.5 m on each side (Figure 7.42). The pressure p_o at the outlet of the valve is constant, and it is less than p . The valve resistance is approximately linear, and $R = 1000$ Pa·s/kg. Assume that the process is isothermal.
 - a. Develop a mathematical model of the pressure p in the container.
 - b.  Construct a Simulink block diagram to find the output $p(t)$ of the pneumatic system if the pressure inside the container is initially 2 atm and the pressure at the outlet is 1 atm.
2. Figure 7.43 shows a liquid-level system in which two tanks have hydraulic capacitances C_1 and C_2 , respectively. The volume flow rate into tank 1 is q_i . The liquid flows from tank 1 to tank 2 through a valve of linear resistance R_1 and leaves tank 2 through a valve of linear resistance R_2 . The density ρ of the liquid is constant.
 - a. Derive the differential equations in terms of the liquid heights h_1 and h_2 . Write the equations in matrix form.

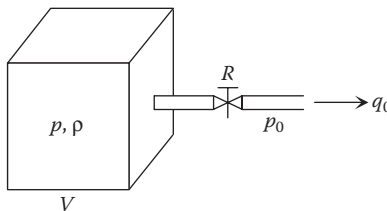


FIGURE 7.42
Problem 1.

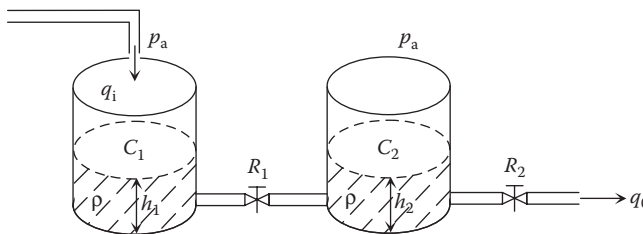


FIGURE 7.43
Problem 2.

- b. Assume that the volume flow rate q_i is the input and the liquid heights h_1 and h_2 are the outputs. Determine the state-space form of the system.
 - c.  Construct a Simulink block diagram to find the outputs $h_1(t)$ and $h_2(t)$ of the liquid-level system. Assume that $\rho = 1000 \text{ kg/m}^3$, $g = 9.81 \text{ m/s}^2$, $C_1 = 0.15 \text{ kg}\cdot\text{m}^2/\text{N}$, $C_2 = 0.25 \text{ kg}\cdot\text{m}^2/\text{N}$, $R_1 = R_2 = 100 \text{ N}\cdot\text{s}/(\text{kg}\cdot\text{m}^2)$, and initial liquid heights $h_1(0) = 1 \text{ m}$ and $h_2(0) = 0 \text{ m}$. The volume flow rate q_i is a step function with a magnitude of 0 before $t = 0 \text{ s}$ and a magnitude of $0.25 \text{ m}^3/\text{s}$ after $t = 0 \text{ s}$.
3. A chicken is taken out of the oven at a uniform temperature of 150°C and is left out in the open air at the room temperature of 25°C . Assume that the chicken can be approximated as a lumped model. The estimated parameters are mass $m = 1.75 \text{ kg}$, heat transfer surface area $A = 0.3 \text{ m}^2$, specific heat capacity $c = 3220 \text{ J}/(\text{kg}\cdot^\circ\text{C})$, and heat transfer coefficient $h = 15 \text{ W}/(\text{m}^2\cdot^\circ\text{C})$.
- a. Derive the differential equation relating the chicken's temperature $T(t)$ and the room temperature.
 - b.  Using the differential equation obtained in Part (a), construct a Simulink block diagram and find the temperature of the chicken.
 - c.  Build a Simscape model of the system, and find the temperature output of the chicken.
 - d. Assume that the chicken can be served only if its temperature is higher than 80°C . Based on the simulation results obtained in Parts (b) and (c), can the chicken be left at the room temperature of 25°C for 1 hour?

7.5 Summary

This chapter was devoted to the modeling of fluid and thermal systems. For each of them, we first introduced the concepts of capacitance and resistance. It is useful to think of fluid and thermal systems as electrical circuits. Along with the basic elements, the conservation of mass and the conservation of energy are the main laws used to develop mathematical models of fluid and thermal systems, respectively.

Fluid systems can be divided into pneumatics and hydraulics. A pneumatic system is one in which the fluid is compressible. At low pressure and moderate or high temperature, real gases may be approximated as ideal gases to simplify calculations for pneumatic systems. A hydraulic system is one in which the fluid is incompressible. Most liquids are generally considered incompressible, and this approximation greatly simplifies the modeling of hydraulic systems. A general category of hydraulic systems is the liquid-level systems.

Fluid capacitance is the relation between the stored fluid mass and the resulting pressure caused by the stored mass,

$$C = \frac{dm}{dp}$$

The pneumatic capacitance of a container of constant volume V is defined as

$$C = \frac{V}{nR_g T}$$

where the value of n depends on the type of the thermodynamic process. For a tank of cross-sectional area A , with a liquid of height h , the hydraulic capacitance of the tank is defined as

$$C = \frac{A(h)}{g}$$

When a fluid flows through a valve, a pipe, or an orifice, the fluid meets resistance, and there is a drop in the pressure of the fluid. The pressure difference is associated with the mass flow rate q_m in a nonlinear relationship. Near a reference operating point, linearization can be performed to obtain the linearized resistance,

$$R = \frac{\Delta p}{\Delta q_m}$$

To obtain a simple model of a fluid system, each mass storage element can be represented by a capacitance element, and each valve can be represented by a resistance element. The differential equation of the system can be derived by applying the law of conservation of mass:

$$\frac{dm}{dt} = q_{mi} - q_{mo}$$

where the mass flow rate into or out of the system can be related to the resistance at the inlet or the outlet of the system, respectively. The resulting mathematical model may adequately describe the dynamics of the real system.

A thermal system is one that involves the transfer of heat from one object to another. For an object, the thermal capacitance is defined as the ratio of the change in heat flow to the change in the object's temperature,

$$C = \frac{dQ}{dT}$$

For incompressible liquids and solids,

$$C = mc = \rho Vc$$

where c is the specific heat capacity.

The thermal resistance for heat transfer is defined as the ratio of the change in temperature difference to the change in heat flow rate,

$$R = \frac{dT}{dq_h}$$

For simple one-dimensional conduction, Fourier's law gives

$$q_h = kA \frac{\Delta T}{L} = kA \frac{T_1 - T_2}{L}$$

and the thermal resistance for conduction is

$$R = \frac{L}{kA}$$

For convective heat transfer, Newton's law of cooling gives

$$q_h = hA\Delta T = hA(T_s - T_{\text{env}})$$

and the thermal resistance for convection is

$$R = \frac{1}{hA}$$

The mathematical model of a thermal system is often complicated because of the complex temperature distribution throughout the system. To simplify analysis, a lumped-parameter model may be used to approximate the gross system dynamics. The validity of this lumped-parameter assumption can be checked using the so-called Biot number,

$$Bi = \frac{hL_c}{k}$$

where L_c is the characteristic length of the solid object, defined as

$$L_c = \frac{V_{\text{body}}}{A_{\text{surface}}}$$

For a body whose Biot number is much less than one, that is, $Bi < 0.1$, the interior of the body may be assumed to have a uniform temperature. Then, the dynamic model of a heat transfer system can be derived using the law of conservation of energy,

$$\frac{dU}{dt} = q_{\text{hi}} - q_{\text{ho}}$$

Review Problems

1. Dry air at a constant temperature of T passes through a valve into a rigid spherical container (Figure 7.44). The pressure at the inlet is p_i , and the one at the outlet is p_a . The linear resistances of the two valves at the inlet and the outlet are R_1 and R_2 , respectively. Assume that the process is isothermal. Develop a mathematical model of the pressure p in the container.
2. A single-tank liquid-level system is shown in Figure 7.45, in which water flows into the tank at a volume flow rate q_i and out of the tank through two valves at points 1 and 2. The linear resistances of the two valves are R_1 and R_2 , respectively. Assuming that $h > h_v$, derive the differential equation relating the liquid height h and the volume flow rate q_i at the inlet. The cross-sectional area of the tank A is constant. The density ρ of the liquid is constant.
3. A watermelon is taken out of the refrigerator at a uniform temperature of 3°C and is exposed to 32°C air. Assume that the watermelon can be approximated as a sphere and the temperature of the watermelon is uniform. The estimated

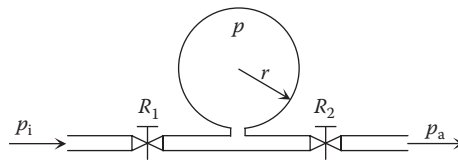


FIGURE 7.44
Problem 1.

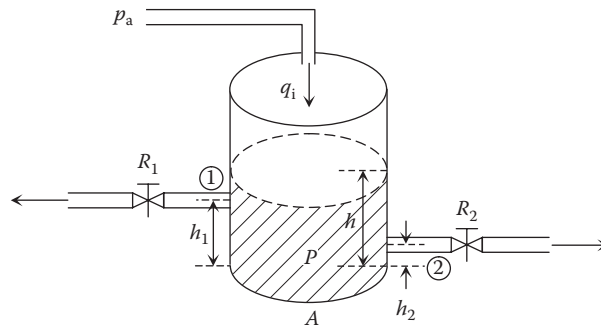


FIGURE 7.45
Problem 2.

parameters are density $\rho = 120 \text{ kg/m}^3$, diameter $D = 35 \text{ cm}$, specific heat capacity $c = 4200 \text{ J/(kg}\cdot^\circ\text{C)}$, and heat transfer coefficient $h = 15 \text{ W/(m}^2\cdot^\circ\text{C)}$.

- Derive the differential equation relating the watermelon's temperature $T(t)$ and the air temperature.
 - Using the differential equation obtained in Part (a), construct a Simulink block diagram and find the temperature of the watermelon.
 - Build a Simscape model of the system.
 - Based on the simulation results obtained in Parts (b) and (c), how long will it take before the watermelon is warmed up to 20°C ?
4. Case study

Figure 7.46 represents the temperature dynamics of two adjacent objects, where the thermal capacitances of the objects are C_1 and C_2 , respectively. Assume that the temperatures of both objects are uniform, and they are T_1 and T_2 , respectively. The heat flow rate into object 1 is q_0 , and the temperature surrounding object 2 is T_0 . There are two modes of heat transfer involved, conduction between the objects

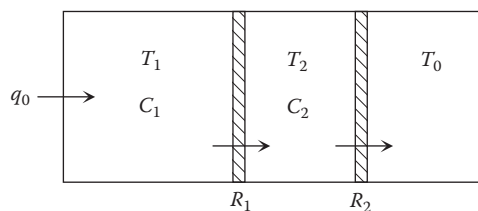


FIGURE 7.46
Problem 4.

and convection between object 2 and the air. The corresponding thermal resistances are R_1 and R_2 , respectively.

- a. Derive the differential equations relating the temperatures T_1 , T_2 , the input q_0 , and the outside temperature T_0 .
- b.  Build a Simscape model of the physical system, and find the temperature outputs $T_1(t)$ and $T_2(t)$. Use default values for the blocks of Thermal Mass (mass = 1 kg, specific heat = 447 J·K/kg, and initial temperature = 300 K), Conductive Heat Transfer (area = 1×10^{-4} m², thickness = 0.1 m, and thermal conductivity = 401 W/(m·K)), and Convective Heat Transfer (area = 1×10^{-4} m² and heat transfer coefficient = 20 W/(m²·K)). Assume that the heat flow rate is $q_0 = 400$ J/s and the surrounding temperature is $T_0 = 298$ K.
- c.  Build a Simulink block diagram based on the differential equations obtained in Part (a), and find the temperature outputs $T_1(t)$ and $T_2(t)$.



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8

System Response

Mathematical models of dynamic systems were presented in [Chapters 5](#) through [7](#), and several forms of system model representations were thoroughly covered in [Chapter 4](#). In this chapter, we present the transient response and frequency response analysis of dynamic systems. Methods to find the solution of the state equation in closed form will also be discussed. For the most part, linear dynamic systems are considered in this chapter. Response of nonlinear systems by using MATLAB® and Simulink® is presented in the final section of the chapter.

8.1 Types of Response

Consider an n th-order dynamic system governed by the differential equation

$$x^{(n)} + a_1 x^{(n-1)} + \dots + a_{n-1} \dot{x} + a_n x = f(t) \quad (8.1)$$

where:

the coefficients $a_1, a_2, \dots, a_n$ are constants

$x(t)$ is the dependent variable

t denotes time

$f(t)$ is the input known as the forcing function

As seen in [Chapter 2](#), the solution $x(t)$, the total response, comprises the complementary (homogeneous) solution $x_c(t)$ and the particular solution $x_p(t)$, that is, $x(t) = x_c(t) + x_p(t)$. Note that $x_c(t)$ is the solution of Equation 8.1 when $f(t) = 0$, and $x_p(t)$ depends on the nature of the forcing function $f(t)$, as well as on its relation to the independent segments of $x_c(t)$. The complementary solution $x_c(t)$ is called the natural response (*free response*), because it represents the natural behavior of the system in the absence of the input. The particular solution $x_p(t)$ is known as the *forced response* of the system.

8.1.1 Transient Response and Steady-State Response

The total response $x(t)$ can also be decomposed into transient response $x_{tr}(t)$ and steady-state response $x_{ss}(t)$, as $x(t) = x_{tr}(t) + x_{ss}(t)$. The *transient response* consists of those terms in $x(t)$ that decay to zero after a sufficiently long time, that is, as $t \rightarrow \infty$. The portion of the response $x(t)$ that remains after the transient terms have vanished is called the *steady-state response*.

Example 8.1: Transient and Steady-State Responses

The model of a dynamic system is derived as

$$3\ddot{x} + 7\dot{x} + 2x = 50\sin t, \quad x(0) = -1, \quad \dot{x}(0) = 1$$

Find the system's total response, and identify the transient and steady-state responses.

Solution

The characteristic values $\lambda = -2, -\frac{1}{3}$ lead to $x_c(t) = c_1 e^{-2t} + c_2 e^{-t/3}$. The particular solution is found by the method of undetermined coefficients ([Chapter 2](#)) as $x_p(t) = -7\cos t - \sin t$. Therefore,

$$x(t) = c_1 e^{-2t} + c_2 e^{-t/3} - 7\cos t - \sin t$$

Initial conditions yield $c_1 = -\frac{12}{5}$, $c_2 = \frac{42}{5}$. The total response is therefore

$$x(t) = -\frac{12}{5} e^{-2t} + \frac{42}{5} e^{-t/3} - 7\cos t - \sin t$$

Consequently, $x_{tr}(t) = -\frac{12}{5} e^{-2t} + \frac{42}{5} e^{-t/3}$ and $x_{ss}(t) = -7\cos t - \sin t$. Note that in this specific example, it turns out that x_c is the transient response and x_p is the steady-state response. This is because the characteristic values are both located in the left half-plane, and the forcing function happens to be sinusoidal.

8.2 Transient Response of First-Order Systems

Linear, first-order dynamic systems are modeled as

$$\tau \dot{x} + x = f(t), \quad \tau = \text{const} > 0, \quad x(0) = x_0 \quad (8.2)$$

where τ is the time constant. Note that upon division by τ , Equation 8.2 agrees with the general form of Equation 8.1. Taking the Laplace transform of Equation 8.2 leads to

$$\tau[sX(s) - x_0] + X(s) = F(s) \quad \xRightarrow[\text{and rearrange}]{\text{Simplify}} \quad X(s) = \frac{\tau}{\tau s + 1} x_0 + \frac{1}{\tau s + 1} F(s)$$

Subsequently, inverse Laplace transformation yields

$$x(t) = \mathcal{L}^{-1} \left\{ \frac{\tau}{\tau s + 1} \right\} x_0 + \mathcal{L}^{-1} \left\{ \frac{1}{\tau s + 1} F(s) \right\}$$

Noting the inverse Laplace transform in the first term is $e^{-t/\tau}$, the total response is expressed as

$$x(t) = \underbrace{\boxed{e^{-t/\tau} x_0}}_{\text{Zero-input response}} + \underbrace{\mathcal{L}^{-1} \left\{ \frac{1}{\tau s + 1} F(s) \right\}}_{\text{Zero-state response}} \quad (8.3)$$

The first term on the right side of Equation 8.3 represents the system response to initial condition (initial excitation) only, and is called the zero-input response. The second term describes the response to the input, and is known as the zero-state response. In this section, we will study the response of first-order systems subjected to specific types of input such as step and ramp functions. The system's free response is discussed first.

8.2.1 Free Response of First-Order Systems

Free response is defined as the response to the initial condition only, hence given by the first term in Equation 8.3,

$$x(t) = e^{-t/\tau} x_0 \quad (8.4)$$

It is readily seen that the smaller the time constant, the faster the response reaches equilibrium.

Example 8.2: Free Response

Suppose a first-order system is described by

$$4\dot{x} + 3x = 0, \quad x(0) = \frac{3}{4}$$

The response $x(t)$ is clearly a free (natural) response as there is no forcing function present. Rewriting the ODE as $\frac{4}{3}\dot{x} + x = 0$, the time constant is identified as $\tau = \frac{4}{3}$. With this, and noting $x(0) = \frac{3}{4}$, the free response of the system is given by Equation 8.4, as

$$x(t) = \frac{3}{4} e^{-3t/4}$$

8.2.2 Impulse Response of First-Order Systems

Impulse response refers to $x(t)$ in Equation 8.2 when $f(t) = A\delta(t)$, where $\delta(t)$ is the unit impulse and A is a constant magnitude (Section 2.3). The Laplace transform of $f(t) = A\delta(t)$ is $F(s) = A$. Inserting this in Equation 8.3, we find

$$x(t) = e^{-t/\tau} x_0 + \mathcal{L}^{-1} \left\{ \frac{A}{\tau s + 1} \right\}$$

which simplifies to

$$x(t) = e^{-t/\tau} x_0 + \frac{A}{\tau} e^{-t/\tau} \quad (8.5)$$

Since $\tau > 0$, we have $x(t) \rightarrow 0$ as $t \rightarrow \infty$. That is, the impulse response has a steady-state value of 0.

8.2.3 Step Response of First-Order Systems

Step response refers to $x(t)$ in Equation 8.2 when $f(t) = Au(t)$, where $u(t)$ is the unit step and A is a constant magnitude (Section 2.3). The Laplace transform of $f(t) = Au(t)$ is $F(s) = A/s$. Inserting this into Equation 8.3, we find

$$x(t) = e^{-t/\tau} x_0 + \mathcal{L}^{-1} \left\{ \frac{A}{s(\tau s + 1)} \right\}$$

Using partial-fraction expansion, it can be shown that

$$\mathcal{L}^{-1} \left\{ \frac{A}{s(\tau s + 1)} \right\} = A(1 - e^{-t/\tau})$$

Therefore, the step response is described by

$$x(t) = e^{-t/\tau} x_0 + A(1 - e^{-t/\tau}) \quad (8.6)$$

Since $\tau > 0$, we have $x(t) \rightarrow A$ as $t \rightarrow \infty$. In other words, the step response has a steady-state value of A .

Example 8.3: Step Response of an RL Circuit

Consider the RL circuit shown in Figure 8.1a, consisting of a resistance R and an inductance L . The initial current is $i(0) = i_0$. Assuming the applied voltage v_a is modeled as a step function with magnitude V , find the zero-input, zero-state, and steady-state responses.

Solution

Using Kirchhoff's voltage law (KVL), the circuit is modeled as

$$L \frac{di}{dt} + Ri = v_a(t), \quad i(0) = i_0 \quad \xRightarrow{\text{Divide by } R} \quad \frac{L}{R} \frac{di}{dt} + i = \frac{1}{R} v_a(t), \quad i(0) = i_0$$

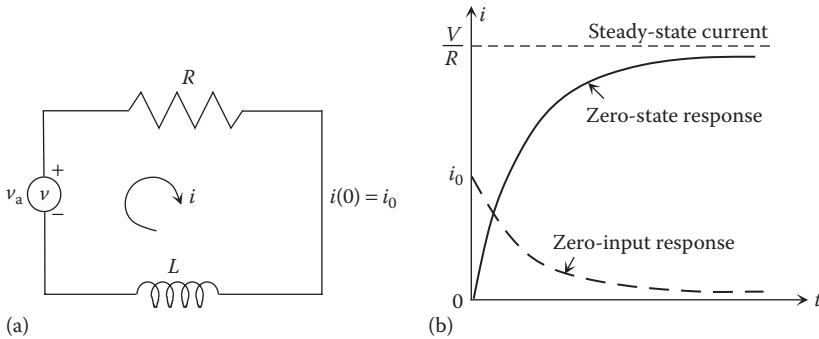


FIGURE 8.1

(a) RL circuit and (b) response to a step input.

Therefore, the time constant is identified as $\tau = L/R$. Comparing this last equation with Equation 8.2, the forcing function is $f(t) = (1/R)v_a(t)$. But since $v_a(t) = Vu(t)$, this forcing function is written as $f(t) = (V/R)u(t)$. Therefore, we can interpret $f(t)$ as a step function with magnitude V/R . As a result, using $A = V/R$ and $x_0 = i_0$ in Equation 8.6, we have

$$i(t) = i_0 e^{-Rt/L} + \frac{V}{R} (1 - e^{-Rt/L})$$

The first term describes the zero-input response, while the second term represents the zero-state response. It is clear that $i(t)$ reaches a steady-state value of V/R after a sufficiently long time, [Figure 8.1\(b\)](#), hence the steady-state current is $i_{ss} = V/R$.

The Role of Time Constant in Speed of Response

We saw earlier that the smaller the time constant, the faster the response reaches steady-state. We will elaborate on this as follows. Consider the RL circuit in Example 8.3, and assume zero initial current, $i(0) = 0$. As a result, the current is given by

$$i(t) = \frac{V}{R} (1 - e^{-t/\tau})$$

where we have replaced L/R with the generic notation τ for the time constant. After one time constant ($t = \tau$), the current is calculated as

$$i(\tau) = \frac{V}{R} (1 - e^{-1}) = 0.632 \left(\frac{V}{R} \right) = 0.632 i_{ss}$$

This means 63.2% of the steady-state current is recovered after one time constant. Similarly, the percentage recovery at $t = 2\tau$ is 86.5% and at $t = 3\tau$ is 95%. At $t = 4\tau$, we have

$$i(4\tau) = \frac{V}{R} (1 - e^{-4}) = 0.982 \left(\frac{V}{R} \right) = 0.982 i_{ss}$$

Therefore, after four time constants, the response is within 2% of the steady-state value; see [Figure 8.2](#). This essentially serves as the settling time for the step response curve. As we discover in [Chapter 10](#), settling time is one of the four transient-response specifications of second-order systems and plays a central role in the control of such systems.

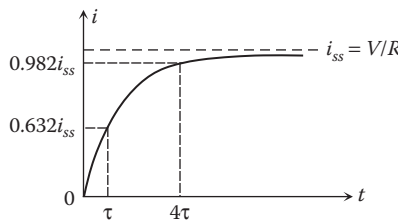


FIGURE 8.2

Role of time constant in speed of response.

8.2.4 Ramp Response of First-Order Systems

Ramp response refers to response $x(t)$ in Equation 8.2 when the forcing function is $f(t) = Au_r(t)$, where $u_r(t)$ is the unit ramp and A is a constant slope (Section 2.3). Inserting $F(s) = A/s^2$ in Equation 8.3, we find

$$x(t) = e^{-t/\tau}x_0 + \mathcal{L}^{-1}\left\{\frac{A}{s^2(\tau s + 1)}\right\}$$

Noting that

$$\mathcal{L}^{-1}\left\{\frac{A}{s^2(\tau s + 1)}\right\} = A\left[t - \tau(1 - e^{-t/\tau})\right]$$

the ramp response is expressed as

$$x(t) = e^{-t/\tau}x_0 + A\left[t - \tau(1 - e^{-t/\tau})\right] \quad (8.7)$$

8.2.4.1 Steady-State Error

For simplicity, suppose the initial condition is zero so that Equation 8.7 simplifies to

$$x(t) = A\left[t - \tau(1 - e^{-t/\tau})\right]$$

The error between the ramp input and the ramp response is

$$e(t) = \underbrace{At}_{\text{Ramp input}} - \underbrace{A\left[t - \tau(1 - e^{-t/\tau})\right]}_{\text{Ramp response}} = A\tau(1 - e^{-t/\tau})$$

As $t \rightarrow \infty$, this error approaches $A\tau$; thus, the steady-state error is

$$e_{ss} = A\tau$$

The ramp input, ramp response and the corresponding steady-state error are shown in Figure 8.3.

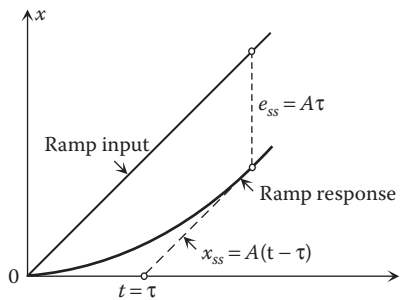


FIGURE 8.3

Ramp response of a first-order system.

PROBLEM SET 8.1

- Find the transient and steady-state responses of a system modeled as $2\ddot{x} + 3\dot{x} = e^{-t}$, $x(0) = 0$, $\dot{x}(0) = 1$.
- Consider a first-order system with time constant τ and zero initial condition. Find the system's unit-impulse response for $\tau = \frac{1}{4}$ and $\tau = \frac{3}{4}$, plot the two curves versus $0 \leq t \leq 2$ in the same graph, and comment.
- Consider a first-order system with time constant τ and zero initial condition. Find the system's unit-step response for $\tau = \frac{1}{3}$ and $\frac{2}{3}$, plot the two curves versus $0 \leq t \leq 2$ in the same graph, and comment.
- The temperature of a steel sphere with initial temperature T_0 submerged in water with temperature T_i is governed by the initial-value problem

$$RC\dot{T} + T = T_i, \quad T(0) = T_0$$

where:

R is the thermal resistance due to convection

C is the thermal capacitance of the sphere

Find the temperature of the sphere at any time, as well as its steady-state value.

- Repeat Problem 4 for the case when the temperature of the water tank increases linearly with time at a rate of r .
- A single-tank liquid-level system with inflow rate q_i as its input and liquid level h as its output is modeled as $RA\dot{h} + gh = Rq_i(t)$, $h(0) = 0$, where $R, A, g = \text{const}$. If the inflow rate is a unit step, find the liquid level in terms of the physical parameters. Also find the steady-state value of the liquid level.
- The equation of motion of the mechanical system in Figure 8.4 is $b\dot{y} + k(y - x) = 0$ where x and y are the input and the output, respectively, and $b, k = \text{const}$. Assuming zero initial condition, find the response when x is a
 - Unit impulse
 - Unit step
- The equation of motion for the torsional mechanical system in Figure 8.5 is derived as $J\ddot{\theta} + B\dot{\theta} = T(t)$, where $J, B = \text{const}$, θ is the angular displacement, and T is a constant applied torque. Express the model as first-order in angular velocity $\omega = \dot{\theta}$. Assuming $\omega(0) = \omega_0$, find $\omega(t)$. Also identify the transient and steady-state responses.

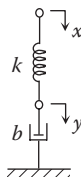


FIGURE 8.4
Problem 7.

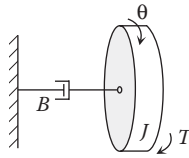


FIGURE 8.5
Problem 8.

9. The governing equation for an RC circuit driven by an applied voltage $v(t)$ is derived as

$$RC\dot{V} + V = v(t), \quad V(0) = V_0$$

where:

V is the voltage across the capacitor

V_0 is the capacitor voltage at the initial time

Assume that the applied voltage is a ramp with slope of $\frac{1}{2}$. Find the steady-state error between the input and the system response.

10. Find the unit-ramp response of the RL circuit in Example 8.3.
11. A first-order dynamic system is modeled as

$$\dot{y} + 3y = f(t), \quad y(0) = 1$$

Assuming the input $f(t)$ is a step function with magnitude 0.8, find y_{ss} .

12. A first-order dynamic system is modeled as

$$\frac{1}{2}\dot{w} + 5w = g(t), \quad w(0) = \frac{2}{3}$$

If the input $g(t)$ is a ramp function with a slope of $\frac{5}{2}$, find the steady-state response and the steady-state error.

8.3 Transient Response of Second-Order Systems

Linear, second-order dynamic systems are mathematically modeled as

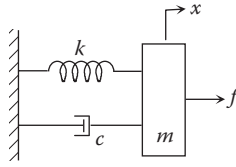
$$\ddot{x} + 2\zeta\omega_n\dot{x} + \omega_n^2x = f(t), \quad x(0) = x_0, \quad \dot{x}(0) = \dot{x}_0 \quad (8.8)$$

where:

ζ is the damping ratio

ω_n is the (undamped) natural frequency, in radians/second

Although Equation 8.8 represents the model for any second-order dynamic system, it is best understood when it is viewed in relation to a mechanical system. To that end, consider the mass-spring-damper system in [Figure 8.6](#).

**FIGURE 8.6**

A mass–spring–damper system.

The system's (undamped) natural frequency is defined as

$$\omega_n = \sqrt{\frac{k}{m}} \text{ rad/sec}$$

The damping ratio is defined as the ratio of the actual damping c and the critical damping*:

$$\zeta = \frac{c}{c_{cr}} = \frac{c}{2\sqrt{mk}}$$

As shown in [Chapter 5](#), the equation of motion is derived as

$$m\ddot{x} + c\dot{x} + kx = f(t) \quad \xRightarrow{\text{Divide by } m} \quad \ddot{x} + \frac{c}{m}\dot{x} + \frac{k}{m}x = \frac{1}{m}f(t)$$

Noting that

$$\omega_n^2 = \frac{k}{m}, \quad 2\zeta\omega_n = 2\frac{c}{2\sqrt{mk}}\sqrt{\frac{k}{m}} = \frac{c}{m}$$

the equation of motion can be expressed in the form of Equation 8.8, with the force $(1/m)f(t)$ re-named $f(t)$. The transient-response analysis of second-order systems is performed as follows. Laplace transformation of Equation 8.8, taking into account the initial conditions, yields

$$\left[s^2X(s) - sx_0 - \dot{x}_0 \right] + 2\zeta\omega_n \left[sX(s) - x_0 \right] + \omega_n^2X(s) = F(s)$$

Collecting like terms and solving for $X(s)$, we find

$$X(s) = \frac{(s + 2\zeta\omega_n)x_0 + \dot{x}_0}{s^2 + 2\zeta\omega_ns + \omega_n^2} + \frac{1}{s^2 + 2\zeta\omega_ns + \omega_n^2}F(s)$$

Finally,

$$x(t) = \underbrace{\mathcal{L}^{-1}\left\{\frac{(s + 2\zeta\omega_n)x_0 + \dot{x}_0}{s^2 + 2\zeta\omega_ns + \omega_n^2}\right\}}_{\text{Zero-input response}} + \underbrace{\mathcal{L}^{-1}\left\{\frac{1}{s^2 + 2\zeta\omega_ns + \omega_n^2}F(s)\right\}}_{\text{Zero-state response}} \quad (8.9)$$

* Critical damping c_{cr} is the value of c that satisfies $c^2 - 4mk = 0$, that is, $c_{cr} = 2\sqrt{mk}$.

The first term on the right side represents the zero-input response, which is the response to initial conditions only. The second term describes the zero-state response, which is the response to the input only.

8.3.1 Free Response of Second-Order Systems

Free response is the response to initial conditions only, defined by the first term in Equation 8.9,

$$x(t) = \mathcal{L}^{-1} \left\{ \frac{(s + 2\zeta\omega_n)x_0 + \dot{x}_0}{s^2 + 2\zeta\omega_n s + \omega_n^2} \right\} \quad (8.10)$$

This inverse Laplace transform depends on the nature of the poles, that is, the roots of the characteristic equation

$$s^2 + 2\zeta\omega_n s + \omega_n^2 = 0$$

Solution of this equation yields the poles

$$s = -\zeta\omega_n \pm \sqrt{(\zeta\omega_n)^2 - \omega_n^2} = -\zeta\omega_n \pm \omega_n\sqrt{\zeta^2 - 1} \quad (8.11)$$

The system is undamped if $\zeta = 0$, underdamped if $0 < \zeta < 1$, critically damped if $\zeta = 1$, and overdamped if $\zeta > 1$. The free response analysis for all these damping cases is presented in the following.

Case (1): Undamped ($\zeta = 0$)

Inserting $\zeta = 0$ in Equation 8.10, we find

$$x(t) = \mathcal{L}^{-1} \left\{ \frac{s x_0 + \dot{x}_0}{s^2 + \omega_n^2} \right\} = x_0 \cos \omega_n t + \frac{\dot{x}_0}{\omega_n} \sin \omega_n t \quad (8.12)$$

Case (2): Underdamped ($0 < \zeta < 1$)

In this case, rewrite

$$\begin{aligned} s^2 + 2\zeta\omega_n s + \omega_n^2 &= (s + \zeta\omega_n)^2 - (\zeta\omega_n)^2 + \omega_n^2 \\ &= (s + \zeta\omega_n)^2 + \omega_n^2(1 - \zeta^2) \\ &= (s + \zeta\omega_n)^2 + \omega_d^2 \end{aligned}$$

where $\omega_d = \omega_n\sqrt{1 - \zeta^2}$ is the (damped) natural frequency. With this, Equation 8.10 yields

$$x(t) = \mathcal{L}^{-1} \left\{ \frac{(s + 2\zeta\omega_n)x_0 + \dot{x}_0}{(s + \zeta\omega_n)^2 + \omega_d^2} \right\} = e^{-\zeta\omega_n t} \left[x_0 \cos \omega_d t + \frac{\zeta\omega_n x_0 + \dot{x}_0}{\omega_d} \sin \omega_d t \right] \quad (8.13)$$

Case (3): Critically damped ($\zeta = 1$)

Using $\zeta = 1$ in Equation 8.10 leads to

$$x(t) = \mathcal{L}^{-1} \left\{ \frac{(s + 2\omega_n)x_0 + \dot{x}_0}{(s + \omega_n)^2} \right\} = e^{-\omega_n t} [x_0 + (\omega_n x_0 + \dot{x}_0)t] \quad (8.14)$$

Case (4): Overdamped ($\zeta > 1$)

In this case, the two poles are real and distinct as in Equation 8.11. Let

$$s_1 = -\zeta\omega_n + \omega_n\sqrt{\zeta^2 - 1}, \quad s_2 = -\zeta\omega_n - \omega_n\sqrt{\zeta^2 - 1}$$

Then

$$x(t) = \mathcal{L}^{-1} \left\{ \frac{(s + 2\zeta\omega_n)x_0 + \dot{x}_0}{s^2 + 2\zeta\omega_n s + \omega_n^2} \right\} = \frac{-s_2 x_0 + \dot{x}_0}{s_1 - s_2} e^{s_1 t} - \frac{-s_1 x_0 + \dot{x}_0}{s_1 - s_2} e^{s_2 t} \quad (8.15)$$

Note that besides the response for the undamped case, which is oscillatory, all other responses stabilize at zero. The response for the underdamped case exhibits decaying oscillations.

8.3.1.1 Free Response in MATLAB

The `initial` command finds the free response of a state-space model with an initial condition on the states:

$$\begin{cases} \dot{\mathbf{x}} = \mathbf{A}\mathbf{x}, & \mathbf{x}(0) = \mathbf{x}_0 \\ \mathbf{y} = \mathbf{C}\mathbf{x} \end{cases}$$

Then, `initial(sys,x0)` plots the response of system `sys` to an initial condition vector `x0`. The duration of simulation is determined automatically to reflect adequately the response transients. If it is desired to save the output data for future use, then type `[out, t]=initial(sys,x0)`. The output is now saved under `out` and may be plotted via `plot(t,out)`.

Example 8.4: Free Response

A second-order dynamic system is modeled as

$$4\ddot{x} + 4\dot{x} + 5x = 0, \quad x(0) = 0.4, \quad \dot{x}(0) = 0$$

- Find the free response $x(t)$.
-  Plot the free response by using the `initial` command.

Solution

- Rewriting in standard form, we have $\ddot{x} + \dot{x} + \frac{5}{4}x = 0$. Comparing with Equation 8.8,

$$\begin{cases} \omega_n^2 = \frac{5}{4} \\ 2\zeta\omega_n = 1 \end{cases} \Rightarrow \begin{cases} \omega_n = \frac{\sqrt{5}}{2} \text{ rad/sec} \\ \zeta = \frac{1}{\sqrt{5}} < 1 \text{ (underdamped)} \end{cases}$$

Since the system is underdamped, we also calculate $\omega_d = \omega_n\sqrt{1 - \zeta^2} = 1 \text{ rad/sec}$. Using these and the given initial conditions in Equation 8.13, we find $x(t) = \frac{1}{5}e^{-t/2}(2\cos t + \sin t)$.

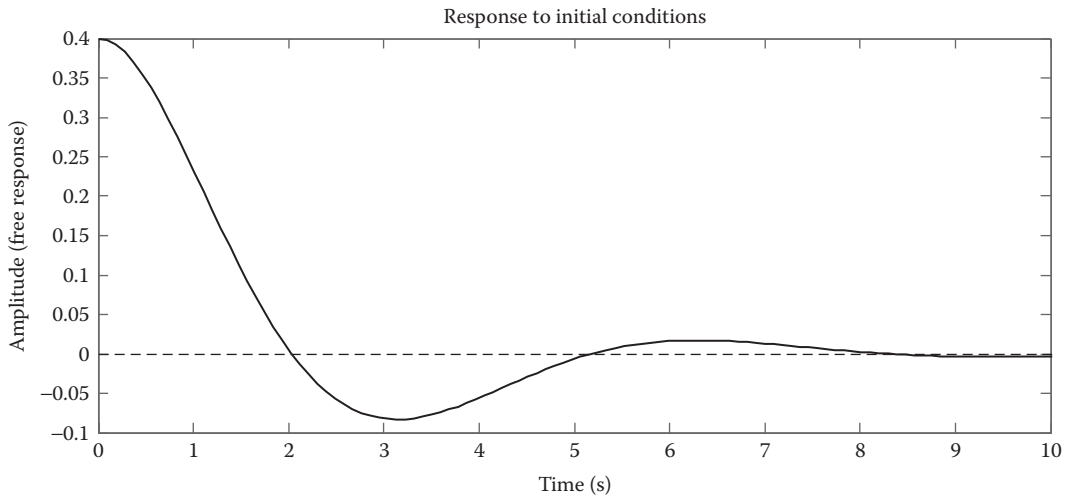


FIGURE 8.7
Free response (Example 8.4).

- b.  Choosing the state variables as $x_1 = x$ and $x_2 = \dot{x}$, the state-variable equations are

$$\begin{cases} \dot{x}_1 = x_2 \\ \dot{x}_2 = \frac{1}{4}(-4x_2 - 5x_1) \end{cases} \text{ so that } \mathbf{A} = \begin{bmatrix} 0 & 1 \\ -\frac{5}{4} & -1 \end{bmatrix}$$

Since the free response is represented by x , which happens to be the first state x_1 , the output equation is $y = x_1$ so that $\mathbf{C} = [1 \ 0]$. The following script will generate the plot of x versus t .

```
>> A = [0 1; -5/4 -1]; C = [1 0]; x0 = [0.4; 0];
>> sys = ss(A, [], C, []);
>> initial(sys, x0) %Figure 8.7
```

Note that $t = 10(\text{sec})$ has been automatically decided by the `initial` command. The result clearly agrees with that in (a).

8.3.2 Impulse Response of Second-Order Systems

Impulse response refers to $x(t)$ in Equation 8.8 when the input is $f(t) = A\delta(t)$, where $\delta(t)$ is the unit impulse and A is a constant. Inserting $F(s) = A$ in Equation 8.9, the response is given by

$$x(t) = \mathcal{L}^{-1} \left\{ \frac{(s + 2\zeta\omega_n)x_0 + \dot{x}_0 + A}{s^2 + 2\zeta\omega_n s + \omega_n^2} \right\} \quad (8.16)$$

As in free response analysis, different damping cases must be studied separately.

Case (1): Undamped ($\zeta = 0$)

$$x(t) = x_0 \cos \omega_n t + \frac{\dot{x}_0 + A}{\omega_n} \sin \omega_n t \quad (8.17)$$

Case (2): Underdamped ($0 < \zeta < 1$)

$$x(t) = e^{-\zeta\omega_n t} \left[x_0 \cos \omega_d t + \frac{\zeta\omega_n x_0 + \dot{x}_0 + A}{\omega_d} \sin \omega_d t \right] \quad (8.18)$$

Case (3): Critically damped ($\zeta = 1$)

$$x(t) = e^{-\omega_n t} [x_0 + (\omega_n x_0 + \dot{x}_0 + A)t] \quad (8.19)$$

Case (4): Overdamped ($\zeta > 1$)

$$x(t) = \frac{-s_2 x_0 + \dot{x}_0 + A}{s_1 - s_2} e^{s_1 t} - \frac{-s_1 x_0 + \dot{x}_0 + A}{s_1 - s_2} e^{s_2 t} \quad (8.20)$$

where:

$$s_1 = -\zeta\omega_n + \omega_n \sqrt{\zeta^2 - 1}, \quad s_2 = -\zeta\omega_n - \omega_n \sqrt{\zeta^2 - 1}$$

As in the case of free response, the impulse response of the undamped system is oscillatory. In each of the other cases where damping is present, the response will stabilize at zero, while the underdamped response exhibits decaying oscillations.

8.3.2.1 Impulse Response in MATLAB

Zero initial conditions: The command `impulse(sys)` returns the unit-impulse response of the linear, time-invariant (LTI) model `sys` created with either transfer function (TF) or state-space (ss) function, with the assumption of zero initial conditions. For multi-input models, independent impulse commands are applied to each input channel. The time range and number of points are chosen automatically in MATLAB. If it is desired to save the output (named `out`) for future use, simply type `[out,t]=impulse(sys)`.

Non-zero initial conditions: Equations 8.17 through 8.20 provide impulse responses of second-order systems with different damping levels. These equations are generated by Equation 8.9, which gives the response as the sum of the zero-input and zero-state responses. The zero-input response is clearly handled by the `initial` command in MATLAB. Since the input is an impulse, the zero-state response is handled by the `impulse` command. Therefore, the total response is simply the superposition of the outputs generated by the `initial` and `impulse` commands; see Problem Set 8.2.

Example 8.5: Impulse Response, Zero Initial Conditions

A mechanical system model is derived as

$$\ddot{x} + 2\dot{x} + kx = 10\delta(t), \quad x(0) = 0, \quad \dot{x}(0) = 0$$

where k represents the coefficient of stiffness.

- Find the response $x(t)$ associated with $k = 2$ and $k = 4$.
-  Plot the responses associated with $k = 2, 4$ in the same graph by using the `impulse` command.

Solution

a. Comparing with Equation 8.8, we have

$$\begin{cases} \omega_n^2 = k \\ 2\zeta\omega_n = 2 \end{cases} \Rightarrow \begin{cases} \omega_n = \sqrt{k} \text{ rad/sec} \\ \zeta = \frac{1}{\sqrt{k}} < 1 \text{ (underdamped, } k=2,4) \end{cases}$$

Since the system is underdamped in both cases, we calculate $\omega_d = \omega_n \sqrt{1 - \zeta^2} = \sqrt{k-1}$ for each. By Equation 8.18, the two responses are then found as

$$\begin{aligned} k=2, \omega_d=1 & \quad x_1(t) = 3e^{-t} \sin t \\ k=4, \omega_d=\sqrt{3} & \quad x_2(t) = \frac{10}{\sqrt{3}} e^{-t} \sin(t\sqrt{3}) \end{aligned}$$

b.  Since the initial conditions are zero, the impulse command is used. We will create our system by using the transfer function. Because the impulse command returns the unit-impulse response, we must define the transfer function as

$$\frac{10}{s^2 + 2s + k} \quad (k=2,4)$$

The following script will generate and plot the impulse responses:

```
>> n = 10; d1 = [1 2 2]; d2 = [1 2 4];
>> sys1 = tf(n,d1); sys2 = tf(n,d2);
>> impulse(sys1,sys2) %Figure 8.8
```

Since the damping ratio and stiffness are related via $\zeta = 1/\sqrt{k}$, the damping ratio associated with $k=2$ is larger than that for $k=4$.

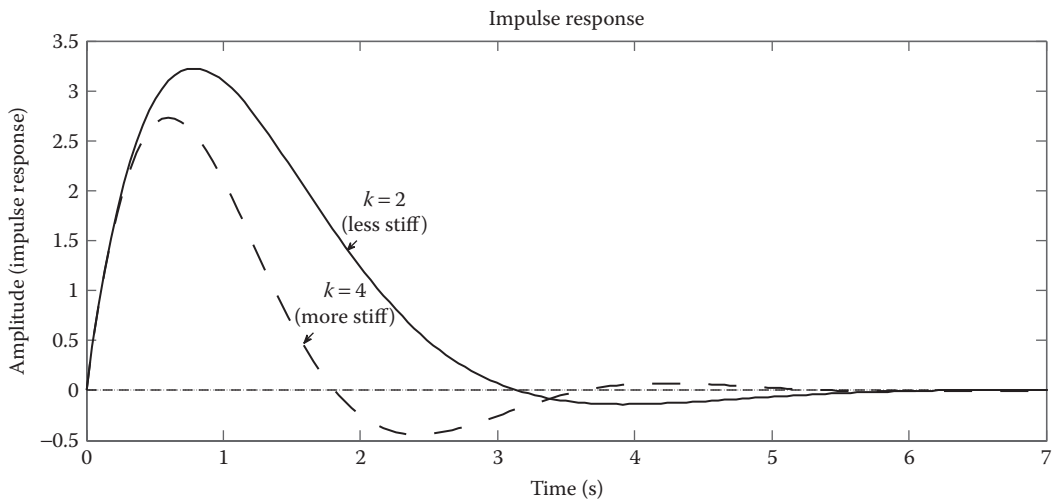


FIGURE 8.8
Impulse responses (Example 8.5).

8.3.3 Step Response of Second-Order Systems

Step response refers to $x(t)$ in Equation 8.8 when the input is $f(t) = Au(t)$, where $u(t)$ is the unit step and A is a constant. Using $F(s) = A/s$ in Equation 8.9, the response is obtained as

$$x(t) = \mathcal{L}^{-1} \left\{ \frac{(s + 2\zeta\omega_n)x_0 + \dot{x}_0}{s^2 + 2\zeta\omega_n s + \omega_n^2} \right\} + \mathcal{L}^{-1} \left\{ \frac{A}{s(s^2 + 2\zeta\omega_n s + \omega_n^2)} \right\} \quad (8.21)$$

As always, different cases of damping are considered separately as follows.

Case (1): Undamped ($\zeta = 0$)

$$x(t) = x_0 \cos \omega_n t + \frac{\dot{x}_0}{\omega_n} \sin \omega_n t + \frac{A}{\omega_n^2} (1 - \cos \omega_n t) \quad (8.22)$$

Case (2): Underdamped ($0 < \zeta < 1$)

$$x(t) = e^{-\zeta\omega_n t} \left[x_0 \cos \omega_d t + \frac{\zeta\omega_n x_0 + \dot{x}_0}{\omega_d} \sin \omega_d t \right] + \frac{A}{\omega_n^2} \left[1 - e^{-\zeta\omega_n t} \left(\cos \omega_d t + \frac{\zeta}{\sqrt{1-\zeta^2}} \sin \omega_d t \right) \right] \quad (8.23)$$

Case (3): Critically damped ($\zeta = 1$)

$$x(t) = e^{-\omega_n t} [x_0 + (\omega_n x_0 + \dot{x}_0)t] + \frac{A}{\omega_n^2} [1 - e^{-\omega_n t} (1 + \omega_n t)] \quad (8.24)$$

Case (4): Overdamped ($\zeta > 1$)

$$x(t) = \frac{-s_2 x_0 + \dot{x}_0}{s_1 - s_2} e^{s_1 t} - \frac{-s_1 x_0 + \dot{x}_0}{s_1 - s_2} e^{s_2 t} + \frac{A}{\omega_n^2} \left[1 + \frac{\omega_n^2}{s_1 - s_2} \left(\frac{1}{s_1} e^{s_1 t} - \frac{1}{s_2} e^{s_2 t} \right) \right] \quad (8.25)$$

where:

$$s_1 = -\zeta\omega_n + \omega_n \sqrt{\zeta^2 - 1}, \quad s_2 = -\zeta\omega_n - \omega_n \sqrt{\zeta^2 - 1}$$

Besides the undamped case where the response oscillates, it is readily seen that the step response for each of the damping cases has a steady-state value of A/ω_n^2 .

8.3.3.1 Step Response in MATLAB

Zero initial conditions: The command `step(sys)` returns the unit-step response of the LTI model `sys` created with either transfer function (TF) or state-space (ss) function, with the assumption of zero initial conditions. For multi-input models, independent step commands are applied to each input channel. The time range and number of points are chosen automatically in MATLAB. If it is desired to save the output (named `out`) for future use, simply type `[out,t]=step(sys)`.

Non-zero initial conditions: Equations 8.22 through 8.25 provide step responses of second-order systems with different damping levels. These equations are generated by Equation 8.9, which gives the response as the sum of the zero-input and zero-state responses. The zero-input response is handled by the `initial` command in MATLAB. Since the input is a step, the zero-state response is handled by the `step` command. The total response is simply the superposition of the outputs generated by the `initial` and `step` commands.

Example 8.6: Step Response, Zero Initial Conditions

The mechanical system in [Figure 8.9](#) is subjected to zero initial displacement and velocity.

- Find the response $x(t)$.
- Plot the response by using the `step` command.

Solution

- Using the given physical parameter values, the equation of motion is written as

$$2\ddot{x} + 2\dot{x} + 3x = 20u(t), \quad x(0) = 0, \quad \dot{x}(0) = 0$$

Dividing by 2 and comparing with the standard form of Equation 8.8, we find

$$\begin{cases} \omega_n^2 = \frac{3}{2} \\ 2\zeta\omega_n = 1 \end{cases} \Rightarrow \begin{cases} \omega_n = \sqrt{\frac{3}{2}} \text{ rad/sec} \\ \zeta = \frac{1}{\sqrt{6}} < 1 \end{cases}$$

This implies the system is underdamped and $\omega_d = \omega_n \sqrt{1 - \zeta^2} = \frac{\sqrt{5}}{2} \text{ rad/sec}$. Using Equation 8.23, with zero initial conditions and $A = 10$, we find the response as (in meters)

$$x(t) = \frac{10}{\frac{3}{2}} \left[1 - e^{-t/2} \left(\cos \frac{\sqrt{5}}{2} t + \frac{1}{\sqrt{5}} \sin \frac{\sqrt{5}}{2} t \right) \right]$$

Note that the steady-state value of response is $\frac{A}{\omega_n^2} = \frac{20}{3}$.

- Since the `step` command returns the unit-step response, we will define the transfer function as

$$\frac{20}{2s^2 + 2s + 3}$$

Subsequently, the following script will generate the desired plot.

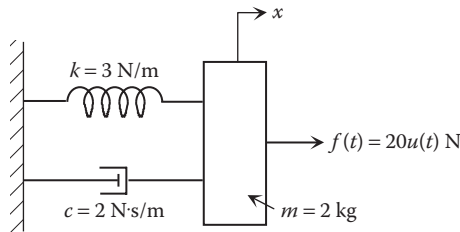


FIGURE 8.9
Mechanical system (Example 8.6).

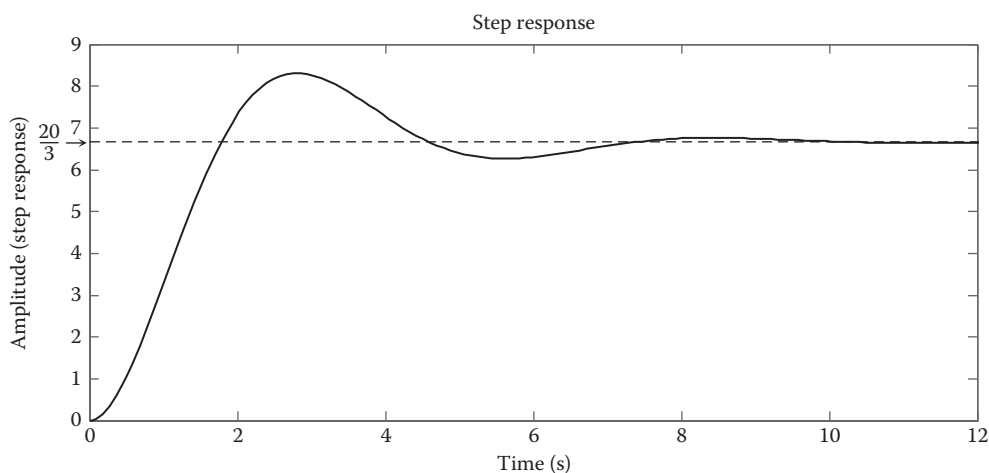


FIGURE 8.10
Step response of the mechanical system in Example 8.6.

```
>> n = 20; d = [2 2 3]; sys = tf(n,d);
>> step(sys)           %Figure 8.10
```

Note that $t = 12(\text{sec})$ has been automatically determined by the `step` command, and that the step response has a steady-state value of $\frac{20}{3}$, as asserted in (a).

Figure 8.11 gives a summary of the behavior of unit-impulse and unit-step responses of second-order systems subjected to zero initial conditions for all three cases of damping. Note that the oscillatory response curves associated with the undamped cases have been omitted.

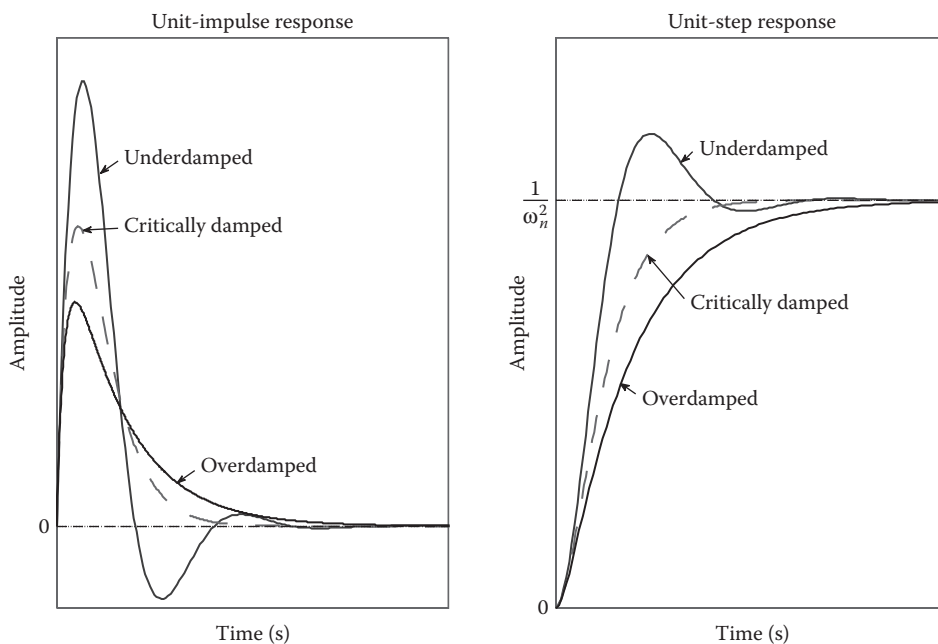


FIGURE 8.11
Unit-impulse and unit-step responses of second-order systems.

8.3.4 Response Analysis Using Simulink

The MATLAB built-in function `initial` returns the system response to initial excitations only, while `impulse` and `step` return responses to impulse and step functions with assumption of zero initial conditions. These tasks may also be achieved through simulation of the Simulink model of the system. This approach proves particularly valuable in applications in which the input is not one of the aforementioned special functions and the initial conditions are non-zero. These general cases can also be handled by the built-in command `lsim`.

8.3.4.1 The `lsim` Command

The `lsim` command, with function call `lsim(sys, u, t, x0)`, plots the time response of the LTI model `sys` to the arbitrary input signal described by `u` and `t`. The time vector `t` consists of regularly-spaced time samples and `u` is a matrix with as many columns as inputs and whose i th row specifies the input value at time $t(i)$. For state-space models only, `x0` is the initial state vector at $t(1)$. It is set to zero when omitted.

Example 8.7: Response Using `lsim` and Simulink

A dynamic system is modeled as

$$4\ddot{x} + 4\dot{x} + 5x = 10\sin(\tfrac{1}{2}t), \quad x(0) = \tfrac{1}{2}, \quad \dot{x}(0) = 0$$

Plot the response $x(t)$ for $0 \leq t \leq 20$ by

- Using the `lsim` command.
- Simulating the Simulink model.

Solution

- With state variables $x_1 = x$, $x_2 = \dot{x}$, we first obtain the state-space model as

$$\begin{cases} \dot{\mathbf{x}} = \begin{bmatrix} 0 & 1 \\ -\frac{5}{4} & -1 \end{bmatrix} \mathbf{x} + \begin{bmatrix} 0 \\ \frac{5}{2} \end{bmatrix} u, \quad \mathbf{x} = \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix}, \quad u = \sin(\tfrac{1}{2}t) \\ y = \begin{bmatrix} 1 & 0 \end{bmatrix} \mathbf{x} \end{cases}$$

Note that the output matrix is chosen as $\mathbf{C} = \begin{bmatrix} 1 & 0 \end{bmatrix}$ since $x = x_1$ is to be plotted. The following script will generate the desired plot.

```
>> A = [0 1; -5/4 -1]; B = [0; 5/2]; C = [1 0];
>> sys = ss(A,B,C,[1]); x0 = [1/2; 0];
>> t = 0:0.01:20; u = sin(t/2);
>> lsim(sys,u,t,x0) %Figure 8.12
```

The `lsim` command returns the input, as well as the response curve as shown in [Figure 8.12](#).

- The Simulink model shown in [Figure 8.13](#) is built according to the state-space model in (a). Double-clicking on each integrator allows us to enter the initial condition for the output signal of that integrator block. Double-clicking on the

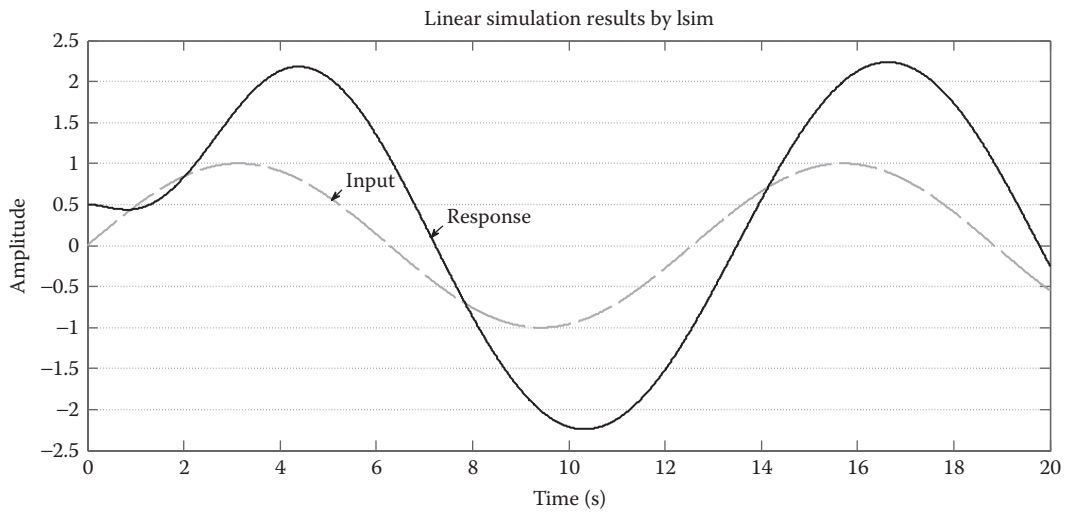


FIGURE 8.12
Simulation result in Example 8.7.

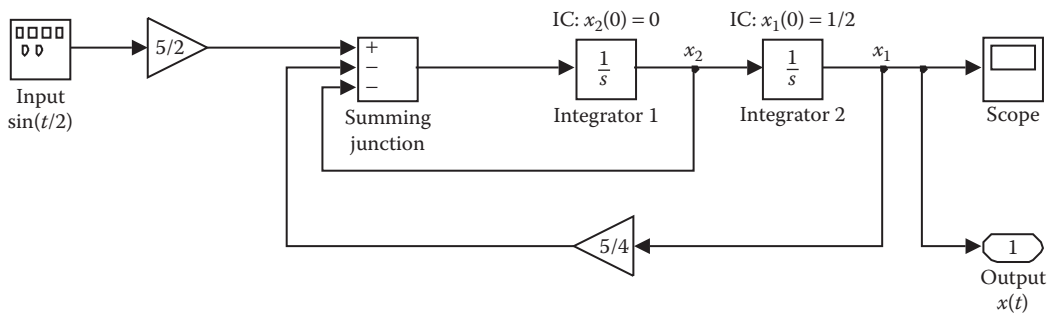


FIGURE 8.13
Simulink model of the system in Example 8.7

input block allows us to select a sine wave with amplitude of 1 and frequency of $\frac{1}{2}$ rad/sec. Running the simulation and double-clicking on the scope block reveals the response plot which agrees with that in Figure 8.12. As always, the plot may be directly accessed by simply typing

```
>> plot(tout,yout)
```

It should be mentioned that the portion of the model that includes the gain of $\frac{5}{2}$ and the two negative feedback loops can be replaced with the State-Space block from the Continuous library. Double-clicking on the block lets us define the proper matrices, as well as the initial conditions; see Example 8.8.

Impulse Response in Simulink

Even though there is no Impulse block in the Sources library of Simulink, the impulse response of a state-space model, subjected to zero initial conditions, may still be obtained using the Simulink model of the system. The model is built as follows. Drag the State-Space block from the Continuous library into the model. Double-clicking on this

block allows us to enter the state-space matrices **A**, **B**, **C**, **D**. For the initial conditions, use the input matrix **B**. Use a Constant block (Commonly Used Blocks) of zero for input. Complete the model by adding the Out and Scope blocks. Running the simulation yields the impulse response of the system. This is illustrated in the following example.

Example 8.8: Impulse Response in Simulink

The model of a dynamic system is derived as $2\ddot{x} + 3\dot{x} + 3x = \delta(t)$ subject to zero initial conditions. Plot $x(t)$ by

- Using the `impulse` command.
- Simulating the system's Simulink model involving the state-space block.

Solution

- Selection of state variables $x_1 = x$, $x_2 = \dot{x}$ leads to the state-space matrices

$$\mathbf{A} = \begin{bmatrix} 0 & 1 \\ -\frac{3}{2} & -\frac{3}{2} \end{bmatrix}, \mathbf{B} = \begin{bmatrix} 0 \\ \frac{1}{2} \end{bmatrix}, \mathbf{C} = \begin{bmatrix} 1 & 0 \end{bmatrix}, \mathbf{D} = 0$$

```
>> A = [0 1; -3/2 -3/2]; B = [0; 1/2]; C = [1 0];
>> sys = ss(A,B,C,[]);
>> impulse(sys) %Figure 8.14
```

- The model is built as shown in [Figure 8.15](#). Double click on the state-space block and enter the matrices **A**, **B**, **C** and **D** as in (a). Set the initial conditions as $[0; 1/2]$, that is, the input matrix **B**. Running the simulation yields the response shown in [Figure 8.14](#).

So far in this chapter, we have mostly studied single-input-single-output (SISO) systems. The following two examples demonstrate how multiple-input-multiple-output (MIMO) systems may be handled through the state-space description of the model.

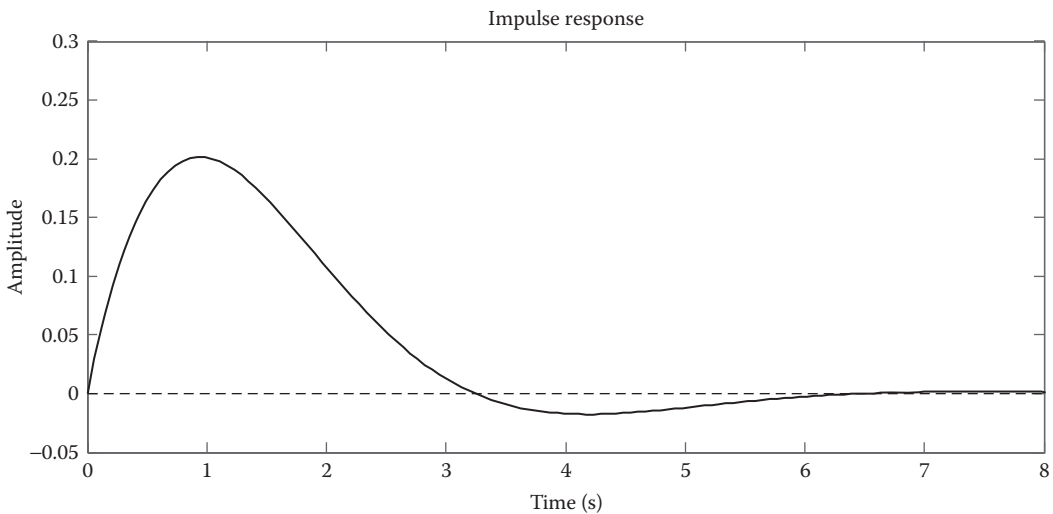


FIGURE 8.14
Impulse response in Example 8.8.

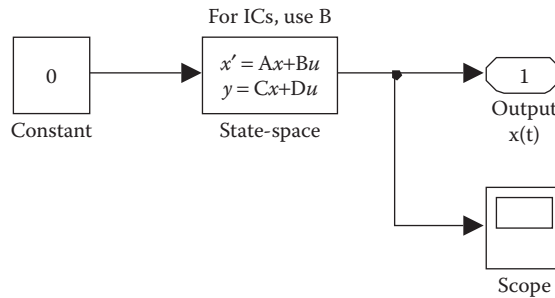


FIGURE 8.15
Simulink model in Example 8.8.

Example 8.9: MIMO System (unit-step input)

In the mechanical system shown in Figure 8.16, all parameter values are in consistent physical units. Assume that the applied forces f_1 and f_2 are unit-step functions and that the system is subject to zero initial conditions. Plot two response curves for x_1 : one for the case when f_1 is the only input, and another for the case when f_2 is the only input. Find the superposition of the two plots to determine the total response x_1 .

Solution

The equations of motion are

$$\begin{cases} 2\ddot{x}_1 + 2x_1 + 0.8\dot{x}_1 - (x_2 - x_1) = f_1 \\ \ddot{x}_2 + x_2 - x_1 = f_2 \end{cases}$$

Selecting state variables $x_1 = x_1$, $x_2 = x_2$, $x_3 = \dot{x}_1$, $x_4 = \dot{x}_2$, the state equation is formed as

$$\dot{\mathbf{x}} = \begin{bmatrix} 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 1 \\ -\frac{3}{2} & \frac{1}{2} & -0.4 & 0 \\ 1 & -1 & 0 & 0 \end{bmatrix} \mathbf{x} + \begin{bmatrix} 0 & 0 \\ 0 & 0 \\ \frac{1}{2} & 0 \\ 0 & 1 \end{bmatrix} \mathbf{u}, \quad \mathbf{x} = \begin{Bmatrix} x_1 \\ x_2 \\ x_3 \\ x_4 \end{Bmatrix}, \quad \mathbf{u} = \begin{Bmatrix} f_1 \\ f_2 \end{Bmatrix}$$

The following script produces a pair of plots, one showing the contribution of f_1 to x_1 (labeled x_{11}), the other reflecting the contribution of f_2 to x_1 (labeled x_{12}). Since the inputs are unit-step functions, and initial conditions are zero, we will use the `step` command in MATLAB.

```
>> A = [0 0 1 0; 0 0 0 1; -3/2 1/2 -0.4 0; 1 -1 0 0];
>> B = [0 0; 0 0; 1/2 0; 0 1];
>> C = [1 0 0 0]; % x1 is to be plotted
```

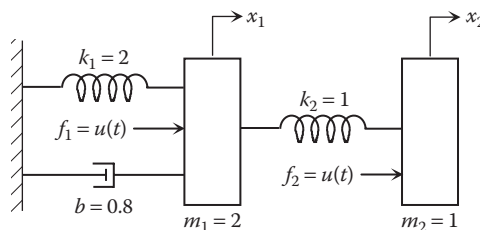


FIGURE 8.16
Mechanical system in Example 8.9.

```

>> sys = ss(A,B,C,[]); % Define system
>> [x1,t] = step(sys); % Results suppressed (see Notes below!)
>> subplot(1,2,1), plot(t,x1(:, :, 1)), % Initiate Figure 8.17a
>> title('Contribution of f1 to x1')
>> subplot(1,2,2), plot(t,x1(:, :, 2)),
>> title('Contribution of f2 to x1') % Complete Figure 8.17a
>> x1 = x1(:, :, 1)+x1(:, :, 2); % Total response by superposition
>> plot(t,x1) %Figure 8.17b

```

Notes

- If only interested in the plots in Figure 8.17a, and not the returned data, simply type

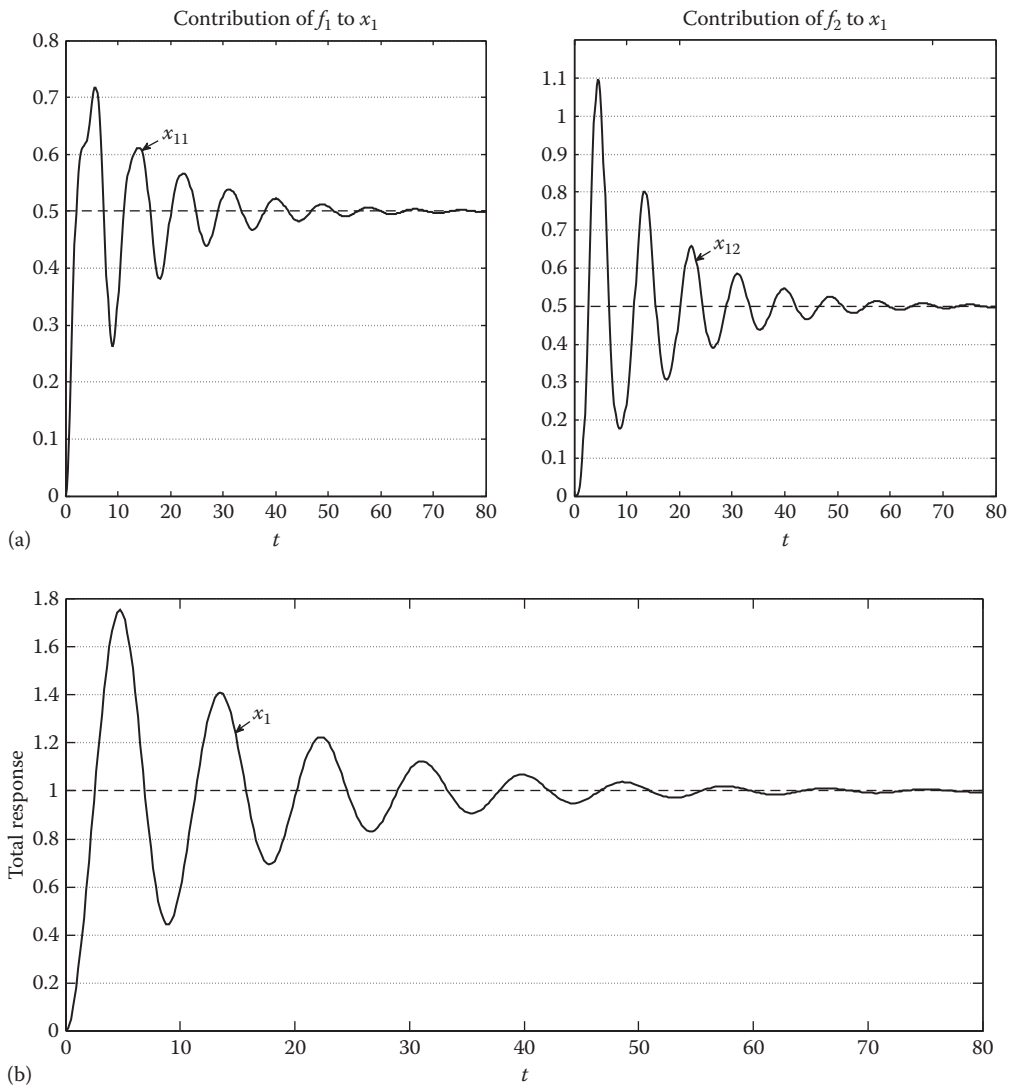


FIGURE 8.17

(a) Contributions of the two inputs to x_1 in Example 8.9 and (b) time variations of response x_1 in Example 8.9.

```
>> step(sys)
```

- The step command automatically determines the suitable time range. A user-specified time range may also be imposed. For example, to generate the plots for up to 100 units of time, type

```
>> step(sys,100)
```

Example 8.10: MIMO System (general input)

Assume the mechanical system in Example 8.9 is subjected to forces $f_1 = e^{-t/2} \sin t$, $f_2 = 0$ and initial conditions $x_1(0) = \frac{1}{2}$, $x_2(0) = 0$, $\dot{x}_1(0) = -1$, $\dot{x}_2(0) = 0$. Plot x_1 versus $0 \leq t \leq 40$.

Solution

```
>> A = [0 0 1 0; 0 0 0 1; -3/2 1/2 -0.4 0; 1 -1 0 0];
>> B1 = [0; 0; 1/2; 0]; % Input f2 is not contributing
>> C = [1 0 0 0]; % x1 is to be plotted
>> t = linspace(0,40,200); % 200 points for a smooth response curve
>> x0 = [1/2; 0; -1; 0]; % Initial state
>> sys = ss(A,B1,C,[]); % Define system
>> u = exp(-t/2).*sin(t); % Define f1
>> lsim(sys,u,t,x0) %Figure 8.18
```

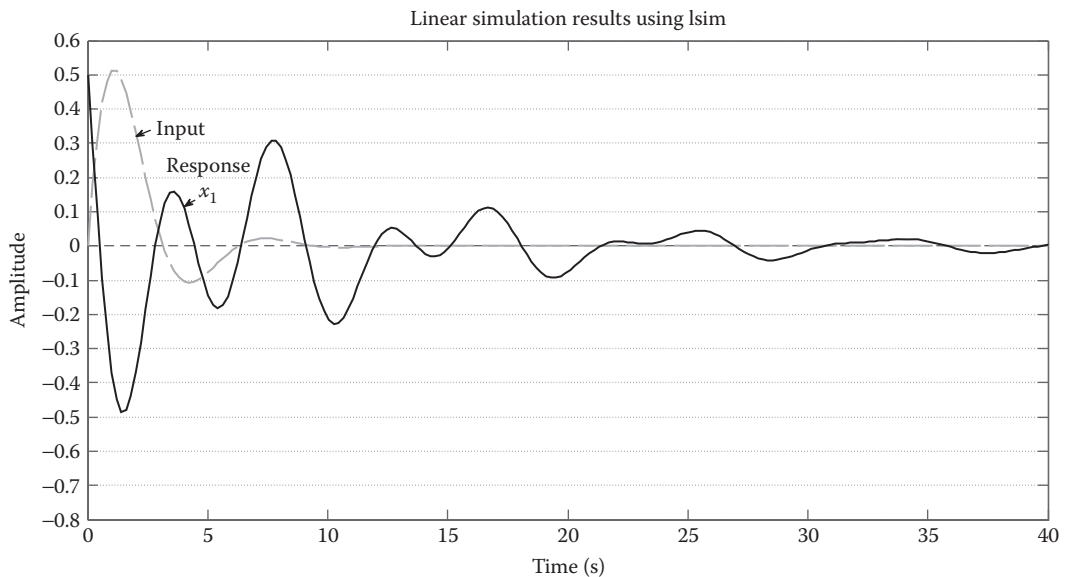


FIGURE 8.18
Response x_1 in Example 8.10.

PROBLEM SET 8.2

1. Show that the free response of an overdamped, second-order system has a steady-state value of zero.

In Problems 2 through 5, for each given system model,

- a. Identify the damping type and find the free response.
 - b.  Plot the free response by using the `initial` command.
2. $3\ddot{x} + 2\dot{x} + x = 0, \quad x(0) = 0, \quad \dot{x}(0) = \frac{1}{3}$
 3. $\ddot{x} + 3\dot{x} + 4x = 0, \quad x(0) = \frac{2}{5}, \quad \dot{x}(0) = 0$
 4. $4\ddot{x} + 8\dot{x} + 3x = 0, \quad x(0) = 1, \quad \dot{x}(0) = -1$
 5. $4\ddot{x} + 4\dot{x} + x = 0, \quad x(0) = \frac{1}{3}, \quad \dot{x}(0) = 1$
 6.  A dynamic system is modeled as

$$4\ddot{x} + c\dot{x} + 5x = 0, \quad x(0) = 1, \quad \dot{x}(0) = 0$$

Plot (in the same graph) the system's free response associated with $c = 4, 5$, and comment.

7. Show that the impulse response of an underdamped, second-order system has a steady-state value of zero.
8. Show that the response of an underdamped, second-order system to an impulsive input $\delta(t - a)$, $a = \text{const} > 0$, and zero initial conditions, is described by

$$\left[\frac{1}{\omega_d} e^{-\zeta\omega_n(t-a)} \sin \omega_d(t-a) \right] u(t-a)$$

where $u(t)$ is the unit-step function. Hint: Use shift on t -axis, [Section 2.3](#). Use this result to find the response $x(t)$ of

$$\ddot{x} + \dot{x} + 0.89x = \delta(t - 1), \quad x(0) = 0, \quad \dot{x}(0) = 0$$

In Problems 9 and 10, assuming zero initial conditions,

- a. Find the response $x(t)$ in closed form.
 - b.  Plot the response by using the `impz` command.
 - c.  Plot the response through the simulation of the Simulink model of the system.
9. $\ddot{x} + 4\dot{x} + x = 5\delta(t)$
 10. $2\ddot{x} + 2\dot{x} + 3x = 0.6\delta(t)$
 11. The governing equation for an RLC circuit driven by the applied voltage $v_a(t)$ is derived as

$$L \frac{di}{dt} + Ri + \frac{1}{C} \int i \, dt = v_a(t)$$

where $L = 4 \text{ H}$, $R = 4 \, \Omega$, and $C = \frac{1}{2} \text{ F}$.

- a. Write the governing equation in terms of the electric charge q .
- b.  Assuming zero initial conditions, plot q and i versus t (same figure) when the applied voltage v_a is a unit impulse.

12. Consider the RLC circuit in Problem 11.
- Write the governing equation in terms of the electric charge q .
 - Assuming initial conditions are $q(0) = 0$, $\dot{q}(0) = 1$, plot q and i versus t (same figure) when the applied voltage v_a is a unit impulse.
13. Consider a single-degree-of-freedom mechanical system as in Figure 8.6 of this section, where $m = 3$ kg, $k = 20$ N/m. Assuming zero initial conditions, plot (in a single figure) the response x to a unit-impulse force for two cases of $c = 2$ N-sec/m and $c = 4$ N-sec/m, and discuss the results.
14. Consider a single-degree-of-freedom mechanical system as in Figure 8.6, where the applied force is $10\delta(t)$ and $m = 9$ kg, $c = 6$ N-sec/m, $k = 5$ N/m. Assume $x(0) = 0.5$, $\dot{x}(0) = 0$.
- Find the response x in closed form.
 - Plot the response x by using the `initial` and `impulse` commands. Also plot the closed-form response of (a) to confirm the result.

In Problems 15 and 16, $u(t)$ denotes the unit step. Assuming zero initial conditions,

- Find the response $x(t)$ in closed form.
 - Plot the response by using the `step` command.
15. $9\ddot{x} + 12\dot{x} + 13x = u(t)$
16. $3\ddot{x} + 12\dot{x} + 10x = 10u(t)$

In Problems 17 through 20, $u(t)$ denotes the unit step.

- Plot the response x by using the `step` and `initial` commands, showing the step response, the initial response and the total response in one figure. Verify the steady-state value of x .
 - Plot the response x by using the `lsim` command.
17. $\frac{1}{4}\ddot{x} + \dot{x} + \frac{5}{4}x = 2u(t)$, $x(0) = 0$, $\dot{x}(0) = \frac{1}{2}$
18. $\ddot{x} + \dot{x} + 4x = 3u(t)$, $x(0) = 0$, $\dot{x}(0) = 1$
19. $9\ddot{x} + 12\dot{x} + 4x = 0.6u(t)$, $x(0) = \frac{2}{3}$, $\dot{x}(0) = 0$
20. $6\ddot{x} + 11\dot{x} + 3x = u(t)$, $x(0) = 1$, $\dot{x}(0) = -\frac{1}{2}$
21. The mechanical system in Figure 8.19, where all parameter values are in consistent physical units, is subject to zero initial conditions. Plot two response curves for x_2 : one corresponding to f_1 as the only input, and another to f_2 only. Finally, plot the total response x_2 .

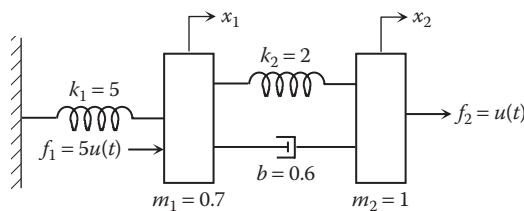


FIGURE 8.19
Problems 21 and 22.

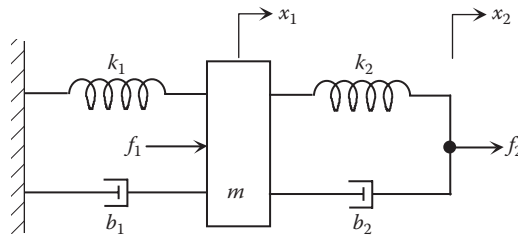


FIGURE 8.20
Problems 24, 25, and 26.

22. The mechanical system in Figure 8.19, where all parameter values are in consistent physical units, is subject to initial conditions $x_1(0) = 1$, $x_2(0) = 1$, $\dot{x}_1(0) = -1$, $\dot{x}_2(0) = 1$. Plot the response x_2 .
23. Reconsider Problem 22. Plot the response x_2 by simulating the Simulink model of the system by using the state-space block.
24. In the mechanical system in Figure 8.20, the parameter values in consistent physical units are

$$m = 1, k_1 = 10, k_2 = 10, b_1 = 5, b_2 = 8$$

The system is subject to initial conditions $x_1(0) = 0$, $x_2(0) = 0$, $\dot{x}_1(0) = -1$ and unit-step applied forces. Plot the response x_1 .

25. Reconsider Problem 24. Plot the response x_1 by simulating the Simulink model of the system by using the state-space block.
26. Reconsider Problem 24 but assume $f_1 = 0$ and f_2 is a unit-impulsive force. All parameter values remain unchanged. Assuming zero initial conditions, plot the response x_2 .

In Problems 27 through 30, the governing equations and initial conditions of a dynamic system are provided. Plot the specified output(s), together with the input(s), by using the `lsim` command. Note that $u(t)$ denotes the unit-step function.

$$27. 8\ddot{x} + 4\dot{x} + 5x = \frac{3}{5}e^{-t/3}\sin t, \quad x(0) = \frac{2}{3}, \quad \dot{x}(0) = 0, \quad 0 \leq t \leq 15$$

Output : $x(t)$

$$28. 64\ddot{x} + 16\dot{x} + 17x = u(t-1), \quad x(0) = 0, \quad \dot{x}(0) = 1, \quad 0 \leq t \leq 40$$

Output : $x(t)$

$$29. \begin{cases} \ddot{x}_1 + \frac{5}{3}x_1 - \frac{2}{3}x_2 - 3(\dot{x}_2 - \dot{x}_1) = u(t) & x_1(0) = 0, \quad \dot{x}_1(0) = 1 \\ \ddot{x}_2 + \frac{2}{3}(x_2 - x_1) + 3(\dot{x}_2 - \dot{x}_1) = \sin t & x_2(0) = 1, \quad \dot{x}_2(0) = 0 \end{cases} \quad 0 \leq t \leq 20$$

Outputs : $x_1(t)$, $x_2(t)$

$$30. \begin{cases} 2\ddot{x}_1 + 2x_1 + 3\dot{x}_1 - \frac{3}{2}(x_2 - x_1) - 4(\dot{x}_2 - \dot{x}_1) = e^{-t/4}\sin t & x_1(0) = 1, \quad \dot{x}_1(0) = 0 \\ \frac{3}{2}(x_2 - x_1) + 4(\dot{x}_2 - \dot{x}_1) = e^{-t/3} & x_2(0) = 1 \end{cases}, \quad 0 \leq t \leq 15$$

Outputs : $x_1(t)$, $x_2(t)$

8.4 Frequency Response

When an LTI system is subjected to a sinusoidal input, characterized by its amplitude and forcing frequency, its response will contain two portions: one that vibrates at the natural frequency of the system, and another that follows the forcing frequency. In the presence of damping, the portion that vibrates at the system's natural frequency will eventually die out, as discussed earlier in this chapter. Therefore, the response at steady-state is sinusoidal and has the same frequency as the input (forcing frequency). The steady-state response to a sinusoidal input is known as the frequency response; [Figure 8.21](#).

8.4.1 Frequency Response of Stable, Linear Systems

Consider a stable, linear time-invariant system described by its transfer function $G(s)$ as in [Figure 8.21](#), with input $f(t)$ and output $x(t)$. Assume that the input is sinusoidal in the form $f(t) = F_0 \sin \omega t$, where F_0 is the amplitude and ω denotes the forcing frequency. The frequency response, denoted by x_{ss} , is the steady-state portion of the response and is determined as follows.

Assume the transfer function is a ratio of two polynomials in s , that is,

$$G(s) = \frac{K(s + z_1)(s + z_2) \cdots (s + z_m)}{(s + p_1)(s + p_2) \cdots (s + p_n)}, \quad K = \text{const} \quad (8.26)$$

Note that since x_{ss} is independent of the initial conditions, they are simply omitted. The transfer function is $G(s) = X(s) / F(s)$; thus, $X(s) = G(s)F(s)$, where $F(s) = F_0 \omega / (s^2 + \omega^2)$. If $G(s)$ has distinct poles only, then partial-fraction expansion yields

$$\begin{aligned} X(s) = G(s)F(s) &= \frac{K(s + z_1)(s + z_2) \cdots (s + z_m)}{(s + p_1)(s + p_2) \cdots (s + p_n)} \cdot \frac{F_0 \omega}{s^2 + \omega^2} \\ &= \frac{c_1}{s + p_1} + \frac{c_2}{s + p_2} + \cdots + \frac{c_n}{s + p_n} + \frac{d}{s + j\omega} + \frac{\bar{d}}{s - j\omega} \end{aligned} \quad (8.27)$$

where:

c_i ($i = 1, 2, \dots, n$) and d are constants

$\bar{d}$ is the complex conjugate of d

Inverse Laplace transformation of $X(s)$ yields

$$x(t) = c_1 e^{-p_1 t} + c_2 e^{-p_2 t} + \cdots + c_n e^{-p_n t} + d e^{-j\omega t} + \bar{d} e^{j\omega t}$$

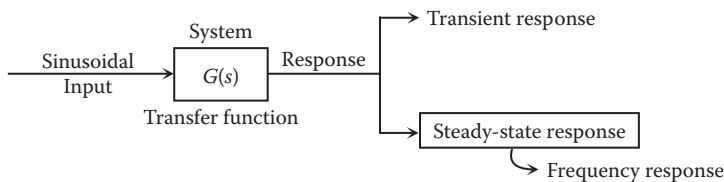


FIGURE 8.21
Frequency response of a linear, time-invariant system.

Because the system is assumed stable, the poles $-p_1, -p_2, \dots, -p_n$ all have negative real parts. Therefore, except for the last two, all terms on the right side die out at steady state. The same result holds for the case of one or more multiple poles. If a typical pole $-p_k$ has multiplicity q , then $x(t)$ will contain terms $t^r e^{-p_k t}$ ($r = 0, 1, \dots, q-1$) so that they too will die out at steady state. Therefore, regardless of the multiplicity of the poles, we have

$$x_{ss}(t) = de^{-j\omega t} + \bar{d}e^{j\omega t} \quad (8.28)$$

The constants d and $\bar{d}$ are evaluated from Equation 8.27 via

$$d = (s + j\omega)X(s)\Big|_{s=-j\omega} = \left[(s + j\omega)G(s) \frac{F_0\omega}{s^2 + \omega^2} \right]_{s=-j\omega} = -\frac{F_0}{2j}G(-j\omega)$$

$$\bar{d} = (s - j\omega)X(s)\Big|_{s=j\omega} = \left[(s - j\omega)G(s) \frac{F_0\omega}{s^2 + \omega^2} \right]_{s=j\omega} = \frac{F_0}{2j}G(j\omega)$$

These results clearly confirm that $\bar{d}$ and d are complex conjugates. The quantity $G(j\omega)$ is known as the frequency response function (FRF) and is obtained by replacing s with $j\omega$ in $G(s)$. Since $G(j\omega)$ is generally complex, it can be expressed in polar form as

$$G(j\omega) = |G(j\omega)|e^{j\phi}, \quad \phi = \angle G(j\omega) = \tan^{-1} \left\{ \frac{\text{imaginary part of } G(j\omega)}{\text{real part of } G(j\omega)} \right\}$$

where ϕ is measured from the positive real axis and is considered positive in the counter-clockwise direction. Similarly,

$$G(-j\omega) = |G(-j\omega)|e^{-j\phi} = |G(j\omega)|e^{-j\phi}$$

Using these in the expressions for d and $\bar{d}$, and inserting the outcomes into Equation 8.28, we find

$$\begin{aligned} x_{ss}(t) &= -\frac{F_0}{2j}|G(j\omega)|e^{-j\phi}e^{-j\omega t} + \frac{F_0}{2j}|G(j\omega)|e^{j\phi}e^{j\omega t} \\ &= F_0|G(j\omega)|\frac{e^{j(\omega t + \phi)} - e^{-j(\omega t + \phi)}}{2j} = F_0|G(j\omega)|\sin(\omega t + \phi) \end{aligned}$$

In summary, if the input of a stable, linear time-invariant system with transfer function $G(s)$ is $F_0 \sin \omega t$, the system's frequency response is

$$x_{ss}(t) = F_0|G(j\omega)|\sin(\omega t + \phi) \quad (8.29)$$

where $|G(j\omega)|$ and ϕ denote, respectively, the magnitude and phase of the FRF $G(j\omega)$.

8.4.1.1 Frequency Response of First-Order Systems

Linear, first-order systems (Section 8.2) are described by

$$\tau \dot{x} + x = f(t), \quad \tau = \text{time constant} > 0$$

Assuming the system is subjected to $f(t) = F_0 \sin \omega t$, its frequency response is obtained as follows. First, the system's transfer function and the ensuing FRF are found as

$$G(s) = \frac{1}{\tau s + 1} \xRightarrow[\text{Replace } s \text{ with } j\omega]{\text{FRF}} G(j\omega) = \frac{1}{1 + \tau\omega j}$$

The magnitude of the FRF is

$$|G(j\omega)| = \frac{1}{\sqrt{1 + (\tau\omega)^2}}$$

To calculate the phase, however, it is advised that $G(j\omega)$ be expressed in standard rectangular form to determine its location in the complex plane. That is,

$$G(j\omega) = \frac{1}{1 + \tau\omega j} = \frac{1 - \tau\omega j}{1 + (\tau\omega)^2} = \frac{1}{1 + (\tau\omega)^2} - j \frac{\tau\omega}{1 + (\tau\omega)^2}$$

Since $\tau\omega > 0$, this indicates that $G(j\omega)$ is always located in the fourth quadrant. As a result, the calculation of phase is straightforward, and it is given by

$$\phi = \angle G(j\omega) = \tan^{-1} \left\{ \frac{\text{imaginary part of } G(j\omega)}{\text{real part of } G(j\omega)} \right\} = \tan^{-1}(-\tau\omega) = -\tan^{-1}(\tau\omega)$$

Therefore, by Equation 8.29, the frequency response is obtained as

$$x_{ss}(t) = \frac{F_0}{\sqrt{1 + (\tau\omega)^2}} \sin(\omega t - \tan^{-1}(\tau\omega)) \quad (8.30)$$

Example 8.11: Frequency Response of a First-Order System

Find the frequency response corresponding to

$$\frac{1}{4} \dot{x} + 3x = \frac{9}{5} \sin\left(\frac{1}{2}t\right)$$

Solution

Rewrite in standard form as $\frac{1}{12} \dot{x} + x = \frac{9}{15} \sin\left(\frac{1}{2}t\right)$ so that $\tau = \frac{1}{12}$, $F_0 = \frac{9}{15}$, and $\omega = \frac{1}{2}$ rad/sec. By Equation 8.30, the frequency response is

$$x_{ss}(t) = \frac{\frac{9}{15}}{\sqrt{1 + \left(\frac{1}{24}\right)^2}} \sin\left(\frac{1}{2}t - \tan^{-1} \frac{1}{24}\right) = 0.5995 \sin\left(\frac{1}{2}t - 0.0416\right)$$

8.4.1.2 Frequency Response of Second-Order Systems

Frequency-response analysis of second-order systems is best understood when applied to a single-degree-of-freedom mechanical system. To that end, consider the mechanical system shown in [Figure 8.22](#), where x is measured from the static equilibrium position and the applied force is $f(t) = F_0 \sin \omega t$. The frequency response x_{ss} is obtained as follows.

Laplace transform of the equation of motion $m\ddot{x} + b\dot{x} + kx = f(t)$, assuming zero initial conditions, leads to the system's transfer function as

$$G(s) = \frac{X(s)}{F(s)} = \frac{1}{ms^2 + bs + k} \xRightarrow{\text{FRF}} G(j\omega) = \frac{1}{k - m\omega^2 + b\omega j}$$

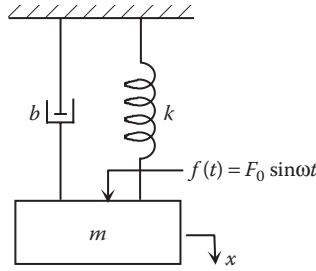


FIGURE 8.22
Second-order mechanical system.

As previously mentioned, calculation of phase requires that we write $G(j\omega)$ in rectangular form to decide its location in the complex plane. That is,

$$G(j\omega) = \frac{k - m\omega^2}{(k - m\omega^2)^2 + (b\omega)^2} - \frac{b\omega}{(k - m\omega^2)^2 + (b\omega)^2} j$$

Since $(k - m\omega^2)^2 + (b\omega)^2 > 0$ and $b\omega > 0$, the imaginary part of $G(j\omega)$ is always negative, placing $G(j\omega)$ below the real axis. In particular, $G(j\omega)$ is in the fourth quadrant if $k - m\omega^2 > 0$ and third quadrant if $k - m\omega^2 < 0$. Therefore, it is necessary to perform this analysis for each given situation. The magnitude and phase of the FRF are determined as

$$|G(j\omega)| = \frac{1}{\sqrt{(k - m\omega^2)^2 + (b\omega)^2}}, \quad \phi = \angle G(j\omega) = -\tan^{-1} \frac{b\omega}{k - m\omega^2}$$

where ϕ must be calculated based on the location of $G(j\omega)$. For a fourth quadrant location, the calculation is straightforward and is simply given by the above-mentioned expression for ϕ . For a third quadrant location, the phase needs to be adjusted accordingly. Recall from [Section 8.3](#) that $\omega_n = \sqrt{\frac{k}{m}}$ and $\zeta = b/(2\sqrt{mk})$, which imply

$$\frac{k}{m} = \omega_n^2, \quad \frac{b}{k} = \frac{2\zeta}{\omega_n}$$

Multiply and divide the fractions involved in $|G(j\omega)|$ and ϕ , and use the above-mentioned relations in the resulting expressions, then insert into Equation 8.29 to obtain

$$x_{ss}(t) = \frac{x_{st}}{\sqrt{[1 - (\omega/\omega_n)^2]^2 + (2\zeta\omega/\omega_n)^2}} \sin\left(\omega t - \tan^{-1} \frac{2\zeta\omega/\omega_n}{1 - (\omega/\omega_n)^2}\right) \quad (8.31)$$

where $x_{st} = F_0/k$ is the static deflection and the dimensionless ratio ω/ω_n is the normalized frequency. If we let X denote the amplitude of x_{ss} in Equation 8.31, then

$$X = \frac{x_{st}}{\sqrt{[1 - (\omega/\omega_n)^2]^2 + (2\zeta\omega/\omega_n)^2}} \quad \Rightarrow \quad \frac{X}{x_{st}} = \frac{1}{\sqrt{[1 - (\omega/\omega_n)^2]^2 + (2\zeta\omega/\omega_n)^2}}$$

The dimensionless ratio X/x_{st} is called the normalized amplitude. Further details of the previous findings and, in particular, applications in mechanical vibrations will be presented in [Section 9.2](#).

Example 8.12: Frequency Response of a Second-Order System

The equation of motion of a mechanical system such as that shown in Figure 8.22 is derived as

$$4\ddot{x} + 4\dot{x} + 13x = 2.4\sin 2t$$

where all parameter values are in consistent physical units. Find the frequency response by using

- Equation 8.31.
- Equation 8.29.

Solution

- It is readily seen that $F_0 = 2.4$, $m = 4$, $b = 4$, $k = 13$, $\omega = 2$ rad/sec. Therefore,

$$G(s) = \frac{1}{4s^2 + 4s + 13} \xrightarrow{\text{FRF}} G(j\omega)\big|_{\omega=2} = \frac{1}{-3 + 8j} = \frac{-3 - 8j}{73}$$

which indicates a third quadrant location. We next find

$$x_{st} = \frac{F_0}{k} = \frac{2.4}{13}, \quad \omega_n = \frac{\sqrt{13}}{2} \text{ rad/sec}, \quad \zeta = \frac{1}{\sqrt{13}}, \quad \frac{\omega}{\omega_n} = \frac{4}{\sqrt{13}}, \quad 2\zeta \frac{\omega}{\omega_n} = \frac{8}{13}$$

To calculate the phase, we first find

$$-\tan^{-1} \frac{2\zeta\omega / \omega_n}{1 - (\omega / \omega_n)^2} = -\tan^{-1} \frac{\frac{8}{13}}{1 - (\frac{4}{\sqrt{13}})^2} = \tan^{-1} \frac{8}{3} = 1.2120 \text{ rad}$$

Taking into account the third quadrant location, the *negative phase* has a magnitude of

$$\phi = \pi - 1.2120 = 1.9296$$

The amplitude in Equation 8.31 is

$$\frac{x_{st}}{\sqrt{[1 - (\omega / \omega_n)^2]^2 + (2\zeta\omega / \omega_n)^2}} = \frac{\frac{2.4}{13}}{\sqrt{(1 - \frac{16}{13})^2 + (\frac{8}{13})^2}} = 0.2809$$

Finally, substitution into Equation 8.31 yields the frequency response

$$x_{ss}(t) = 0.2809 \sin(2t - 1.9296)$$

- The FRF found in (a) is

$$G(2j) = \frac{-3 - 8j}{73} \text{ (Third quadrant)}$$

The magnitude is $|G(2j)| = 1/\sqrt{73}$. Due to third quadrant location, the negative phase is found as

$$\tan^{-1} \left\{ \frac{\frac{8}{\sqrt{3}}}{\frac{3}{\sqrt{3}}} \right\} = \tan^{-1} \frac{8}{3} = 1.2120 \text{ rad} \Rightarrow \phi = \pi - 1.2120 = 1.9296$$

Noting that $F_0 = 2.4$, by Equation 8.29 we have

$$x_{ss}(t) = F_0 |G(2j)| \sin(2t + \phi) = \frac{2.4}{\sqrt{73}} \sin(2t - 1.9296) = 0.2809 \sin(2t - 1.9296)$$

This, as anticipated, agrees with the result of (a).

8.4.2 Bode Plot

The FRF of a dynamic system may be represented by a pair of plots, one displaying the magnitude versus frequency, the other the phase angle (in degrees) versus frequency. A very specific presentation of this pair of plots is known as a Bode plot (or diagram). For a given FRF, denoted by $G(j\omega)$, with magnitude $|G(j\omega)|$ and phase angle ϕ , a Bode diagram consists of two plots: a curve of the (base 10) logarithm of $|G(j\omega)|$ and a plot of the phase ϕ , both versus $\log \omega$. It is common to represent the logarithmic magnitude of $G(j\omega)$ by $20 \log |G(j\omega)|$ with the unit of decibel, abbreviated dB. In a Bode diagram, the curves are sketched on semilog paper, using a linear scale for magnitude (in dB) and phase (in degrees) and a logarithmic scale for frequency (in radians/second). The logarithmic presentation described here allows the low- and high-frequency behavior of the FRF to be shown in a single diagram. This is quite valuable because of the significance of low-frequency characteristics of systems in practical situations. Bode diagrams play an important role in control systems design, as discussed in [Section 10.6](#).

8.4.2.1 Bode Plot in MATLAB

The MATLAB command `bode` generates the Bode plot for an LTI system. For a system `sys`, the command

```
>> bode(sys)
```

will plot the Bode diagram, comprised of a magnitude (dB) plot and a phase (degrees) plot. If the numerical values for magnitude and phase calculated by `bode` are desired, we define a range of frequency in the form of vector `w`, and type

```
>> [mag,phase,w] = bode(sys,w)
```

This returns matrices `mag` and `phase` but not a plot. The desired plots may be generated by executing the following script:

```
>> w = linspace(0.01,100,1000); % 1000 points for smoothness of curves
>> [mag,phase,w]=bode(sys,w);
>> for i=1:1000,
m(i) = mag(1,1,i); % Extract the magnitude column
p(i) = phase(1,1,i); % Extract the phase column
end
>> subplot(2,1,1), semilogx(w,20*log10(m)) % Create the magnitude (dB) plot
>> subplot(2,1,2), semilogx(w,p) % Complete Bode diagram
```

8.4.2.2 Bode Plot of First-Order Systems

Consider a first-order system model in its standard form,

$$\tau \dot{x} + x = f(t), \quad \tau > 0$$

so that the *standard, first-order transfer function* is $G(s) = \frac{1}{(\tau s + 1)}$ and the FRF is formed as

$$G(j\omega) = \frac{1}{1 + \tau\omega j}$$

As shown earlier in this section, the magnitude and phase of the FRF are

$$|G(j\omega)| = \frac{1}{\sqrt{1 + (\tau\omega)^2}}, \quad \phi = -\tan^{-1}(\tau\omega)$$

The Bode plot is obtained as follows. By definition, the logarithmic magnitude is

$$20 \log |G(j\omega)| = 20 \log \frac{1}{\sqrt{1 + (\tau\omega)^2}} = 20 \log [1 + (\tau\omega)^2]^{-1/2} = -10 \log [1 + (\tau\omega)^2] \text{ dB}$$

The plots of this quantity and the phase ϕ of the FRF versus $\tau\omega$ (logarithmic scale) will make up the Bode diagram shown in [Figure 8.23](#). Since τ , the time constant, is in seconds and ω is in rad/sec, $\tau\omega$ is dimensionless and treated as normalized frequency. The asymptotic approximations of magnitude (dB) and phase for low- and high-frequency ranges are determined as follows. In a low-frequency range, we have $\tau\omega \ll 1$ so that $(\tau\omega)^2 \ll 1$ and the logarithmic magnitude is approximated by $-10 \log 1$ dB, or 0 dB. The phase is approximated by $-\tan^{-1}(0)$, or 0° . In a high-frequency range, we have $\tau\omega \gg 1$ so that $(\tau\omega)^2 \gg 1$ and the logarithmic magnitude is approximated by $-10 \log (\tau\omega)^2$ dB, or $-20 \log \tau\omega$ dB.

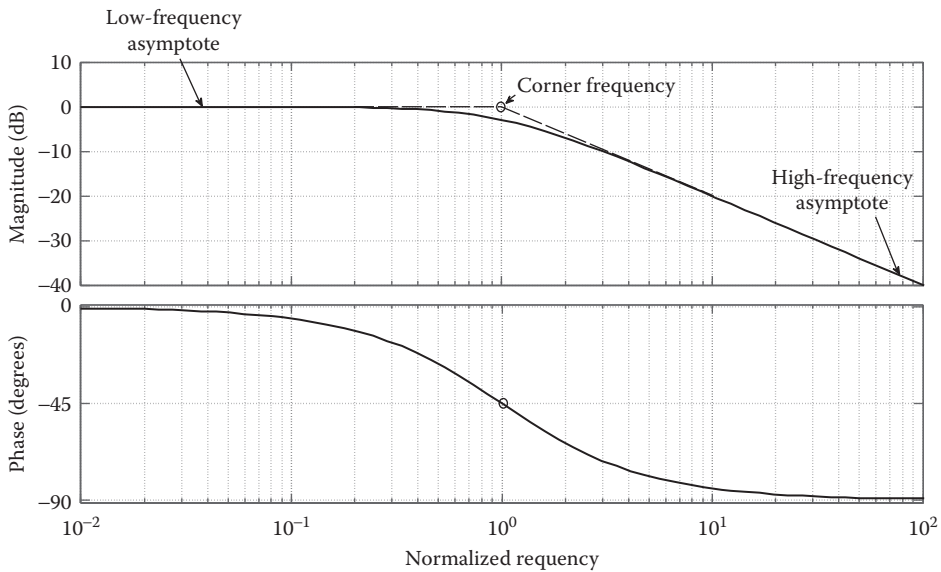


FIGURE 8.23
Bode plot for $G(j\omega) = 1 / (1 + \tau\omega j)$.

This represents a straight line with a slope of -20 dB/decade. The phase will approach $-\tan^{-1}(\infty)$, or -90° . The low- and high-frequency asymptotes intersect when $\tau\omega = 1$, so that $\omega = 1/\tau$ is the *corner frequency*. The phase at the corner frequency is $\phi = -\tan^{-1} 1 = -45^\circ$.

Example 8.13: Bode Plot of a First-Order System

Find the Bode plot and identify the corner frequency:

$$G(s) = \frac{\frac{2}{5}}{3s + 2}$$

Solution

We first rewrite $G(s)$ as

$$G(s) = \frac{\frac{1}{5}}{\frac{3}{2}s + 1} = \frac{1}{5} \cdot \boxed{\frac{1}{\frac{3}{2}s + 1}} \xRightarrow{\text{FRF}} G(j\omega) = \frac{1}{5} \cdot \frac{1}{1 + \frac{3}{2}j\omega}$$

Standard 1st-order TF

This agrees with the standard form $\frac{1}{(1 + \tau\omega j)}$, with $\tau = \frac{3}{2}$, except for the constant multiple of $\frac{1}{5}$. The corner frequency $\omega = \frac{1}{\tau} = \frac{2}{3}$ rad/sec. The magnitude is

$$|G(j\omega)| = \frac{1}{5} \cdot \frac{1}{\sqrt{1 + \left(\frac{3}{2}\omega\right)^2}} \xRightarrow[20\log|G(j\omega)|]{\text{Logarithmic magnitude}} 20\log \frac{1}{5} + 20\log \frac{1}{\sqrt{1 + \left(\frac{3}{2}\omega\right)^2}} \text{ dB}$$

The second term is the standard magnitude (dB) for first-order systems discussed earlier. Since the first term is $20\log \frac{1}{5} = -13.9794$, the complete magnitude (dB) for the system at hand is obtained by lowering the standard magnitude curve by 13.9794 dB. The constant $\frac{1}{5}$ does not affect the phase angle because its phase is zero. The following script generates the desired Bode plot.

```
>> sys = tf(2/5, [3 2]);
>> bode(sys) %Figure 8.24
```

Recall that the same pair of plots can be generated by retaining the numerical values of magnitude and phase using the function call `[mag, phase, w]=bode(sys, w)` and plotting them versus frequency. In the low-frequency range, the asymptote for the standard magnitude is 0 dB as in Figure 8.23. Lowering that by 13.9794 dB yields the result depicted in Figure 8.24. As expected, the phase plot is unchanged compared with the standard case in Figure 8.23.

8.4.2.3 Bode Plot of Second-Order Systems

Consider a second-order system in the standard (normalized) form*

$$\ddot{x} + 2\zeta\omega_n\dot{x} + \omega_n^2x = \omega_n^2f(t) \quad (8.32)$$

so that the *standard, second-order transfer function* is formed as

$$G(s) = \frac{\omega_n^2}{s^2 + 2\zeta\omega_ns + \omega_n^2}$$

* This differs slightly from the form given in Section 8.3. In Equation 8.32, the forcing term is normalized.

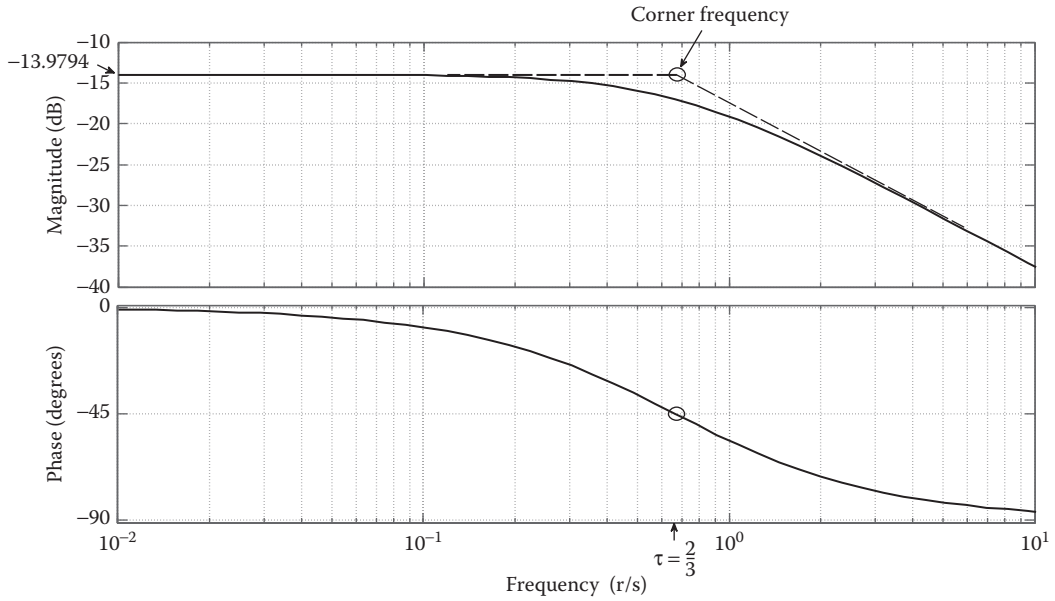


FIGURE 8.24
Bode plot (Example 8.13).

and the FRF is subsequently obtained as

$$G(j\omega) = \frac{\omega_n^2}{\omega_n^2 - \omega^2 + 2\zeta\omega_n\omega j} \quad \text{Divide by } \omega_n^2 \quad = \quad \frac{1}{1 - (\omega / \omega_n)^2 + 2\zeta(\omega / \omega_n)j}$$

The magnitude and phase of the FRF are then calculated as

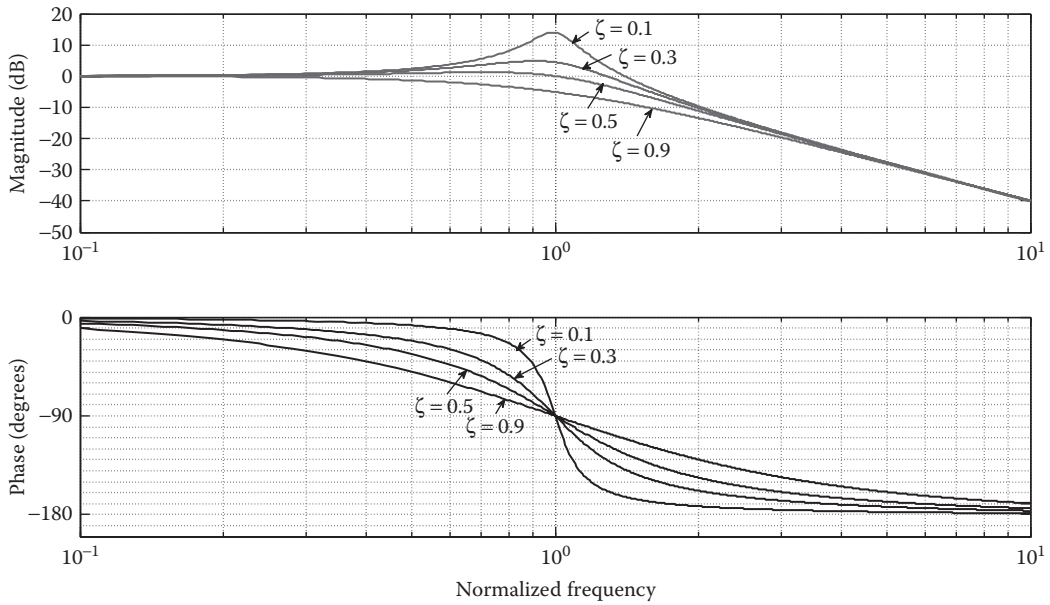
$$|G(j\omega)| = \frac{1}{\sqrt{[1 - (\omega / \omega_n)^2]^2 + (2\zeta\omega / \omega_n)^2}}, \quad \phi = -\tan^{-1} \frac{2\zeta\omega / \omega_n}{1 - (\omega / \omega_n)^2} \quad (8.33)$$

The logarithmic magnitude is

$$20\log|G(j\omega)| = 20\log \frac{1}{\sqrt{[1 - (\omega / \omega_n)^2]^2 + (2\zeta\omega / \omega_n)^2}} \quad \text{dB}$$

For a fixed damping ratio ζ , both the magnitude (dB) and phase angle ϕ are functions of the normalized frequency ω/ω_n and thus may be plotted versus ω/ω_n . Figure 8.25 shows several pairs of curves, each pair corresponding to a specific ζ . It turns out that some magnitude curves experience a maximum peak, while others do not. To find the frequency at which a maximum peak occurs, we proceed as follows. Maximizing the logarithmic magnitude is equivalent to maximizing $|G(j\omega)|$, which has a constant numerator. Therefore, we need to minimize the denominator of $|G(j\omega)|$, that is,

$$\sqrt{(1 - r^2)^2 + (2\zeta r)^2}, \quad r = \frac{\omega}{\omega_n}$$

**FIGURE 8.25**

Bode plot for $G(j\omega) = 1 / [1 - (\omega/\omega_n)^2 + 2\zeta(\omega/\omega_n)j]$.

It is then readily seen that minimization of this quantity with respect to r leads to the critical value

$$r = \frac{\omega}{\omega_n} = \sqrt{1 - 2\zeta^2}$$

With this result, we define the *resonant frequency* ω_r as

$$\omega_r = \omega_n \sqrt{1 - 2\zeta^2} \quad (8.34)$$

Based on this, the magnitude curves that attain a maximum peak correspond to $0 \leq \zeta \leq 0.7071$ where 0.7071 is an approximate value for $\sqrt{2}/2$. The maximum peak of the magnitude itself is then found as

$$|G(j\omega)|_{\max} = |G(j\omega)|_{\omega/\omega_n = \sqrt{1-2\zeta^2}} = \frac{1}{2\zeta\sqrt{1-\zeta^2}} \quad (8.35)$$

For example, the maximum peak for the case of $\zeta = 0.1$ occurs at $\omega/\omega_n = \sqrt{1-2(0.1)^2} = 0.9899$ and

$$|G(j\omega)|_{\max} = \frac{1}{2\zeta\sqrt{1-\zeta^2}} \bigg|_{\zeta=0.1} = 5.0252 \quad \text{Logarithmic magnitude} \Rightarrow 20\log(5.0252) = 14.0231 \text{ dB}$$

All of these are readily verified via [Figure 8.25](#).

Example 8.14: Bode Plot of a Second-Order System

Find the Bode plot for

$$G(s) = \frac{4}{2s^2 + 1.8s + 4.5}$$

Solution

Rewrite $G(s)$ as

$$G(s) = \frac{8}{9} \cdot \frac{2.25}{s^2 + 0.9s + 2.25} \quad \xrightarrow{\text{FRF}} \quad G(j\omega) = \frac{8}{9} \cdot \frac{2.25}{2.25 - \omega^2 + 0.9\omega j}$$

Standard 2nd-order TF

The magnitude is

$$|G(j\omega)| = \frac{8}{9} \cdot \frac{2.25}{\sqrt{(2.25 - \omega^2)^2 + (0.9\omega)^2}} \quad \xrightarrow{\text{Logarithmic magnitude}} \quad 20 \log \frac{8}{9} + 20 \log \frac{2.25}{\sqrt{(2.25 - \omega^2)^2 + (0.9\omega)^2}} \text{ dB}$$

Since the first term is $20 \log \frac{8}{9} = -1.0231$, the complete logarithmic magnitude is obtained by lowering the standard magnitude curve by 1.0231 dB. Once again, the constant $\frac{8}{9}$ does not affect the phase angle. The following script generates the Bode plot in [Figure 8.26](#).

```
>> sys = tf(4, [2 1.8 4.5]);
>> bode(sys) %Figure 8.26
```

The standard, second-order transfer function yields $\omega_n = 1.5$ rad/sec and $\zeta = 0.3$. Since $\zeta \leq 0.7071$, the corresponding magnitude (dB) curve attains a maximum peak, which occurs at the resonant frequency

$$\omega_r = \omega_n \sqrt{1 - 2\zeta^2} \bigg|_{\omega_n=1.5, \zeta=0.3} \cong 1.3583 \text{ rad/sec}$$

The corresponding peak value is calculated as

$$|G(j\omega)|_{\max} = \frac{1}{2\zeta\sqrt{1-\zeta^2}} \bigg|_{\zeta=0.3} = 1.7471 \quad \xrightarrow{\text{Logarithmic magnitude}} \quad 20 \log(1.7471) = 4.8466 \text{ dB}$$

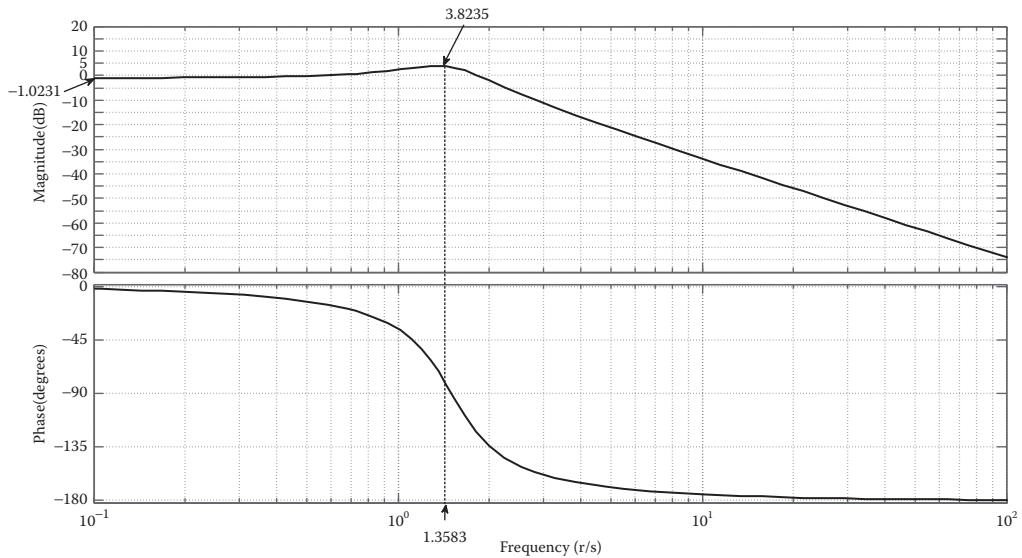


FIGURE 8.26
Bode plot (Example 8.14).

However, since the magnitude curve has been lowered by 1.0231 dB, the maximum peak for the case at hand is adjusted to $4.8466 - 1.0231 = 3.8235$ dB; See [Figure 8.26](#).

Bode Plot of Higher-Order Systems

Bode plot of systems of any order can be constructed using the detailed procedure outlined in [Section 10.6](#). In this section, however, we consider only a specific class of systems whose transfer functions are expressible as

$$G(s) = KG_1(s)G_2(s) \dots G_n(s)$$

where:

- each $G_i(s)$ is either a standard, first-order form or a standard, second-order form
- K is a constant

Let the logarithmic magnitude of each $G_i(s)$ be $20 \log |G_i(j\omega)|$ and its phase ϕ_i . Then, the logarithmic magnitude of $G(s)$ is

$$20 \log |G(j\omega)| = 20 \log |K| + 20 \log |G_1(j\omega)| + \dots + 20 \log |G_n(j\omega)|$$

and since K has a phase of 0, the phase of $G(s)$ is

$$\phi = \phi_1 + \dots + \phi_n$$

Therefore, the magnitude (dB) plot for $G(s)$ is simply obtained by adding the individual magnitude plots of G_i 's and raising (or lowering) the result by $20 \log |K|$. The phase plot of $G(s)$ is obtained by adding the individual phase plots of G_i 's.

Example 8.15: Bode Plot of a Higher-Order System

Draw the Bode plot for

$$G(s) = \frac{5}{(3s+2)(2s^2+1.8s+2.25)}$$

Solution

Rewrite $G(s)$ as

$$G(s) = \frac{5}{(3s+2)(2s^2+1.8s+2.25)} = \boxed{\frac{5}{9}} \cdot \boxed{\frac{1}{\frac{3}{2}s+1}} \cdot \boxed{\frac{2.25}{s^2+0.9s+2.25}}$$

Constant Standard 1st-order Standard 2nd-order

The first-order transfer function is the one involved in Example 8.13, and the second-order in Example 8.14. The constant gain $\frac{5}{9}$ contributes $20 \log \frac{5}{9} = -5.1055$ dB and 0 phase. [Figure 8.27](#) shows the Bode plots for the three individual terms, as well as their sum which represents the Bode plot for the overall transfer function $G(s)$. The following script confirms [Figure 8.27](#).

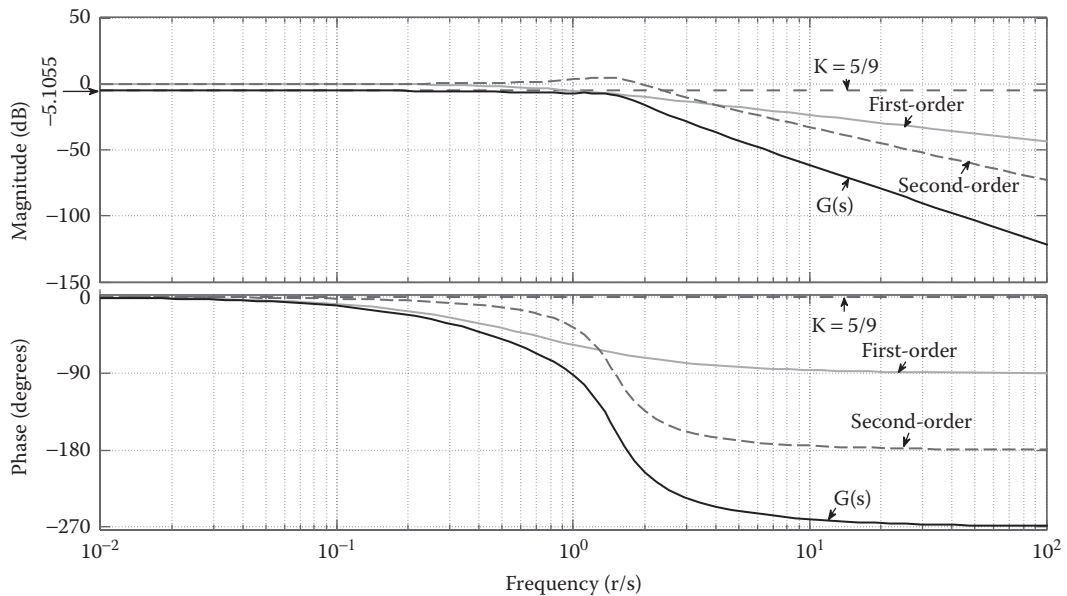


FIGURE 8.27
Composition of Bode plots (Example 8.15).

```
>> sys1 = tf(5/9,1); % Constant gain
>> sys2 = tf(1,[3/2 1]); % Standard, first-order
>> sys3 = tf(2.25,[1 0.9 2.25]); % Standard, second-order
>> sys = tf(5,conv([3 2],[2 1.8 4.5])); % Original transfer function
>> bode(sys1)
>> hold on
>> bode(sys2)
>> bode(sys3)
>> bode(sys) %Figure 8.27
```

PROBLEM SET 8.3

In Problems 1 through 6, find the frequency response of the given system.

1. $\frac{1}{3}\ddot{x} + 2\dot{x} = 2.6 \sin 2t$
2. $\dot{y} + 5y = 14.5 \sin(\frac{1}{2}t)$
3. $4\ddot{x} + 12\dot{x} + 13x = 40 \sin(\frac{1}{3}t)$
4. $2\ddot{x} + 5\dot{x} + 8x = 25 \sin 3t$
5. $4\ddot{x} + 2\dot{x} + 10x = 28.8 \sin(2t)$
6. $10\ddot{x} + 0.8\dot{x} + 20x = 35 \sin t$
7. Consider the mechanical system in [Figure 8.28](#). Assuming $m = 12$ kg, $b = 20$ N-sec/m, $k = 200$ N/m, $F_0 = 47$ N, and $\omega = 5$ rad/sec, find the system's frequency response.
8. Find the frequency response of the undamped mechanical system in [Figure 8.29](#), assuming $m = 10$ kg, $k = 150$ N/m, $F_0 = 115$ N, and $\omega = 4$ rad/sec.

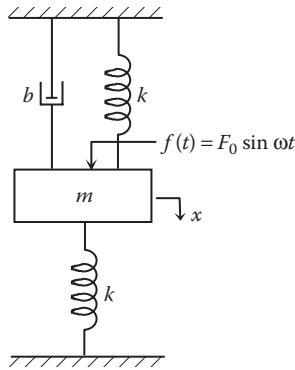


FIGURE 8.28
Problem 7.

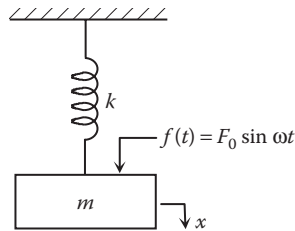


FIGURE 8.29
Problem 8.

9. Show that the frequency response of an RLC circuit driven by a sinusoidal applied voltage $v(t) = V_0 \sin \omega t$ is described by

$$i_{ss}(t) = \frac{\omega C V_0}{\sqrt{(1 - LC\omega^2)^2 + (RC\omega)^2}} \cos\left(\omega t - \tan^{-1} \frac{RC\omega}{1 - LC\omega^2}\right)$$

▲ In Problems 10 through 12, draw the Bode plot and identify the corner frequency, as well as the asymptotic approximations of magnitude for low-frequency and high-frequency ranges.

10. $G(s) = \frac{\frac{2}{5}}{2s + 5}$

11. $G(s) = \frac{4}{3s + \frac{2}{3}}$

12. $G(s) = \frac{2.5}{5s + \frac{3}{4}}$

13. The Bode plot of a first-order system indicates that the corner frequency is 2.25 rad/sec and the approximate low-frequency logarithmic magnitude is -23.10 dB.
- Find the transfer function.
 - Using the results of (a), find the approximate value of the logarithmic gain at $\omega = 100$ rad/sec.

14. Repeat Problem 13 when the corner frequency is 0.45 rad/sec and the approximate low-frequency logarithmic magnitude is 56.25 dB.
15. The Bode plot of a first-order system indicates that the approximate low-frequency and high-frequency (at $\omega = 100$ rad/sec) logarithmic magnitudes are 8.34 dB and -48.53 dB, respectively. Find the system's transfer function.
16. Show that the magnitude $|G(j\omega)| = \frac{1}{\sqrt{[1-(\omega/\omega_n)^2]^2 + (2\zeta\omega/\omega_n)^2}}$ attains a maximum when

$$\frac{\omega}{\omega_n} = \sqrt{1 - 2\zeta^2}$$

▲ In Problems 17 through 19, draw the Bode plot, identify the resonant frequency and the peak magnitude (dB), if applicable, and find the approximate low-frequency logarithmic magnitude.

17. $G(s) = \frac{5}{s^2 + 2s + 9}$
18. $G(s) = \frac{48}{s^2 + 2.6s + 6.76}$
19. $G(s) = \frac{3}{2s^2 + 8.64s + 11.52}$

20. Show that in the Bode plot of the standard, second-order transfer function, the asymptotic approximation of the logarithmic magnitude for the high-frequency range is given by

$$-40 \log \frac{\omega}{\omega_n}$$

which represents a straight line with a slope of -40 dB/decade.

8.5 Analytical Solution of the State Equation

In [Section 4.2](#), we learned that using the state variables, the mathematical model of a linear dynamic system can be expressed in the form of the state equation

$$\dot{\mathbf{x}} = \mathbf{A}\mathbf{x} + \mathbf{B}\mathbf{u}, \quad \mathbf{x}(0) = \mathbf{x}_0 \quad (8.36)$$

where:

- $\mathbf{x}(t)$ is the state vector
- $\mathbf{A}$ is the state matrix
- $\mathbf{B}$ is the input matrix
- $\mathbf{u}(t)$ is the vector of the inputs
- $\mathbf{x}_0$ is the initial state vector

In this section, we will learn how to solve the state equation to derive the state vector in closed form.

8.5.1 Formal Solution of the State Equation

The scalar counterpart of Equation 8.36 is the initial-value problem

$$\dot{x}(t) = ax(t) + bu(t), \quad x(0) = x_0$$

whose formal solution is obtained as

$$x(t) = e^{at} x_0 + \int_0^t e^{a(t-\tau)} bu(\tau) d\tau$$

Inspired by this, the formal solution of Equation 8.36 is expressed as

$$\mathbf{x}(t) = e^{\mathbf{A}t} \mathbf{x}_0 + \int_0^t e^{\mathbf{A}(t-\tau)} \mathbf{B} \mathbf{u}(\tau) d\tau \quad (8.37)$$

The quantity $e^{\mathbf{A}t}$ is known as the matrix exponential of $\mathbf{A}$. Note that technically $e^{\mathbf{A}}$ is the matrix exponential, but $e^{\mathbf{A}t}$ is what is often involved in dynamic systems analysis.

8.5.1.1 Matrix Exponential

When a is scalar, e^{at} is defined as

$$e^{at} = \sum_{k=0}^{\infty} \frac{1}{k!} (at)^k = 1 + ta + \frac{1}{2!} t^2 a^2 + \dots$$

This idea can be extended to the case involving a square matrix $\mathbf{A}_{n \times n}$, as

$$e^{\mathbf{A}t} = \sum_{k=0}^{\infty} \frac{1}{k!} (t\mathbf{A})^k = \mathbf{I} + t\mathbf{A} + \frac{1}{2!} t^2 \mathbf{A}^2 + \dots \quad (8.38)$$

where:

t is a scalar

$\mathbf{I}$ is the $n \times n$ identity matrix and $\mathbf{A}^k = \overbrace{\mathbf{A} \mathbf{A} \dots \mathbf{A}}^{k \text{ times}}$

It is readily seen that the matrix exponential $e^{\mathbf{A}t}$ is $n \times n$.

8.5.1.2 Properties of the Matrix Exponential

1. Given $\mathbf{A}_{n \times n}$ and scalars t and τ ,

$$e^{\mathbf{A}(t+\tau)} = e^{\mathbf{A}t} e^{\mathbf{A}\tau}$$

which can be verified as follows.

$$\begin{aligned}
 e^{\mathbf{A}(t+\tau)} &= \sum_{k=0}^{\infty} \frac{1}{k!} \mathbf{A}^k (t+\tau)^k = \sum_{k=0}^{\infty} \frac{1}{k!} \mathbf{A}^k \sum_{m=0}^{\infty} \binom{k}{m} t^m \tau^{k-m} = \sum_{k=0}^{\infty} \frac{1}{k!} \mathbf{A}^k \sum_{m=0}^{\infty} \frac{k!}{m!(k-m)!} t^m \tau^{k-m} \\
 &= \sum_{m=0}^{\infty} \sum_{k=m}^{\infty} \mathbf{A}^k \frac{t^m \tau^{k-m}}{m!(k-m)!} \underset{k-m=r}{=} \sum_{m=0}^{\infty} \sum_{r=0}^{\infty} \mathbf{A}^{r+m} \frac{t^m \tau^r}{m!r!} \\
 &= \left(\sum_{m=0}^{\infty} \frac{1}{m!} \mathbf{A}^m t^m \right) \left(\sum_{r=0}^{\infty} \frac{1}{r!} \mathbf{A}^r \tau^r \right) = e^{\mathbf{A}t} e^{\mathbf{A}\tau}
 \end{aligned}$$

Here,

$$(t+\tau)^k = \sum_{m=0}^{\infty} \binom{k}{m} t^m \tau^{k-m}$$

is the binomial series, and

$$\binom{k}{m} = \frac{k!}{m!(k-m)!}$$

is the binomial coefficient.

- Since the series in Equation 8.38 is absolutely convergent for finite values of t , term-by-term differentiation of the series yields

$$\frac{d}{dt} e^{\mathbf{A}t} = \mathbf{A} e^{\mathbf{A}t} = e^{\mathbf{A}t} \mathbf{A}$$

- If $\mathbf{A}_{n \times n}$ and $\mathbf{B}_{n \times n}$ are similar matrices such that $\mathbf{S}^{-1} \mathbf{A} \mathbf{S} = \mathbf{B}$, then $\mathbf{A}t = \mathbf{S}[\mathbf{B}t]\mathbf{S}^{-1}$ and

$$e^{\mathbf{A}t} = \mathbf{S} e^{\mathbf{B}t} \mathbf{S}^{-1}$$

- If matrix $\mathbf{D} = [d_{ii}]_{i=1,2,\dots,n}$ is diagonal, then

$$e^{\mathbf{D}t} = [e^{d_{ii}t}]_{i=1,2,\dots,n}$$

Assuming matrix $\mathbf{A}_{n \times n}$ is not defective (that is, it does not have any generalized eigenvectors; see [Section 3.3](#)), the matrix exponential $e^{\mathbf{A}t}$ can be systematically calculated as follows. Suppose $\mathbf{A}_{n \times n}$ has eigenvalues $\lambda_1, \lambda_2, \dots, \lambda_n$ and linearly independent eigenvectors

$\mathbf{v}_1, \mathbf{v}_2, \dots, \mathbf{v}_n$. Then, as shown in [Section 3.3](#), the modal matrix $\mathbf{V} = [\mathbf{v}_1 \ \mathbf{v}_2 \ \dots \ \mathbf{v}_n]$ diagonalizes $\mathbf{A}$ via

$$\mathbf{V}^{-1}\mathbf{A}\mathbf{V} = \mathbf{D} = [\lambda_i]_{i=1,2,\dots,n}$$

Then,

$$\mathbf{A} = \mathbf{V}\mathbf{D}\mathbf{V}^{-1} \xRightarrow{\text{Multiply by scalar } t} \mathbf{A}t = \mathbf{V}[\mathbf{D}t]\mathbf{V}^{-1}$$

Finally, by the above-mentioned property 3, we find

$$e^{\mathbf{A}t} = \mathbf{V}e^{\mathbf{D}t}\mathbf{V}^{-1} \quad (8.39)$$

Noting that the modal matrix $\mathbf{V}$ is available and $e^{\mathbf{D}t}$ is easy to calculate (property 4), $e^{\mathbf{A}t}$ will be readily available.

8.5.1.3 Formal Solution in MATLAB

The command `expm` calculates the matrix exponential.

`expm` Matrix exponential.

`expm(X)` is the matrix exponential of X . `expm` is computed using a scaling and squaring algorithm with a Pade approximation.

Although it is not computed this way, if X has a full set of eigenvectors $\mathbf{V}$ with corresponding eigenvalues $\mathbf{D}$, then $[\mathbf{V}, \mathbf{D}] = \text{EIG}(X)$ and $\text{expm}(X) = \mathbf{V} * \text{diag}(\text{exp}(\text{diag}(\mathbf{D}))) / \mathbf{V}$.

The integration portion of the formal solution, Equation 8.37, is handled by the `int` command.

Example 8.16: Formal Solution

Find the formal solution of

$$\dot{\mathbf{x}} = \begin{bmatrix} 1 & 0 \\ -1 & -2 \end{bmatrix} \mathbf{x} + \begin{bmatrix} 1 \\ 3 \end{bmatrix} e^{-t}, \quad \mathbf{x}_0 = \begin{Bmatrix} 1 \\ 0 \end{Bmatrix}$$

Solution

The formal solution is described by

$$\mathbf{x}(t) = e^{\mathbf{A}t} \mathbf{x}_0 + \int_0^t e^{\mathbf{A}(t-\tau)} \mathbf{B} \mathbf{u}(\tau) d\tau$$

```
>> A = [1 0;-1 -2]; B = [1;3]; x0 = [1;0];
>> syms t tau
>> x = simplify(expm(A*t)*x0+int(expm(A*(t-tau))*B*exp(-tau),tau,0,t))
x =
      (3*exp(t))/2 - exp(-t)/2
- (exp(-2*t)*(exp(3*t) - 7*exp(t) + 6))/2
```

The `simplify` command is used to ensure that the simplified result is returned. The state vector is therefore found as

$$\mathbf{x}(t) = \begin{Bmatrix} \frac{1}{2}(3e^t - e^{-t}) \\ -\frac{1}{2}e^{-2t}(e^{3t} - 7e^t + 6) \end{Bmatrix}$$

8.5.2 Solution of the State Equation via Laplace Transformation

The state equation subjected to a specified initial state vector can also be solved using Laplace transformation. We first note that

$$\mathbf{x} = \begin{Bmatrix} x_1 \\ \dots \\ x_n \end{Bmatrix} \xrightarrow{\text{Laplace transform}} \mathcal{L}\{\mathbf{x}\} = \mathbf{X}(s) = \begin{Bmatrix} \mathcal{L}\{x_1\} \\ \dots \\ \mathcal{L}\{x_n\} \end{Bmatrix}$$

Taking the Laplace transform of Equation 8.36, considering the initial state vector $\mathbf{x}_0$, yields

$$s\mathbf{X}(s) - \mathbf{x}_0 = \mathbf{A}\mathbf{X}(s) + \mathbf{B}\mathbf{U}(s) \xrightarrow{\text{Rearrange}} (s\mathbf{I} - \mathbf{A})\mathbf{X}(s) = \mathbf{x}_0 + \mathbf{B}\mathbf{U}(s)$$

Pre-multiply by $(s\mathbf{I} - \mathbf{A})^{-1}$, known as the *resolvent* of matrix $\mathbf{A}$, to obtain

$$\mathbf{X}(s) = (s\mathbf{I} - \mathbf{A})^{-1}\mathbf{x}_0 + (s\mathbf{I} - \mathbf{A})^{-1}\mathbf{B}\mathbf{U}(s)$$

so that

$$\mathbf{x}(t) = \mathcal{L}^{-1}\{(s\mathbf{I} - \mathbf{A})^{-1}\}\mathbf{x}_0 + \mathcal{L}^{-1}\{(s\mathbf{I} - \mathbf{A})^{-1}\mathbf{B}\mathbf{U}(s)\} \quad (8.40)$$

We will next show that this result exactly agrees with the formal solution, Equation 8.37. The two terms in Equation 8.40 are manipulated as follows. First, recall that for a scalar m ,

$$\frac{1}{1-m} = 1 + m + m^2 + m^3 + \dots \quad (\text{converges for } |m| < 1)$$

Inspired by this, assuming the series converges, we have

$$(\mathbf{I} - \mathbf{M})^{-1} = \mathbf{I} + \mathbf{M} + \mathbf{M}^2 + \dots$$

Then,

$$(s\mathbf{I} - \mathbf{A})^{-1} = \frac{1}{s} \left(\mathbf{I} - \frac{1}{s}\mathbf{A} \right)^{-1} = \frac{1}{s} \left[\mathbf{I} + \frac{1}{s}\mathbf{A} + \frac{1}{s^2}\mathbf{A}^2 + \dots \right] = \frac{1}{s}\mathbf{I} + \frac{1}{s^2}\mathbf{A} + \frac{1}{s^3}\mathbf{A}^2 + \dots$$

Consequently,

$$\mathcal{L}^{-1}\{(s\mathbf{I} - \mathbf{A})^{-1}\} = \mathcal{L}^{-1}\left\{ \frac{1}{s}\mathbf{I} + \frac{1}{s^2}\mathbf{A} + \frac{1}{s^3}\mathbf{A}^2 + \dots \right\} = \mathbf{I} + t\mathbf{A} + \frac{1}{2!}t^2\mathbf{A}^2 + \dots \stackrel{\text{By Equation 8.38}}{=} e^{\mathbf{A}t}$$

This implies the first term in Equation 8.40 is simply $e^{\mathbf{A}t}\mathbf{x}_0$. The second term may be worked out using the convolution integral ([Section 2.3](#)) as

$$\mathcal{L}^{-1}\{(s\mathbf{I} - \mathbf{A})^{-1}\mathbf{B}\mathbf{U}(s)\} = \int_0^t e^{\mathbf{A}(t-\tau)}\mathbf{B}\mathbf{u}(\tau) d\tau$$

Combining the two terms yields

$$\mathbf{x}(t) = e^{\mathbf{A}t}\mathbf{x}_0 + \int_0^t e^{\mathbf{A}(t-\tau)}\mathbf{B}\mathbf{u}(\tau) d\tau$$

which exactly agrees with the formal solution derived earlier.

Example 8.17: Laplace-Transform Approach

Solve the state equation in Example 8.16 by using the Laplace-transform approach.

Solution

The following script will use Equation 8.40 to find the state vector:

```
>> A = [1 0;-1 -2]; B = [1;3]; x0 = [1;0];
>> syms s t
>> U = laplace(exp(-t));
>> x = simplify(ilaplace(inv(s*eye(2)-A))*x0 +
ilaplace(inv(s*eye(2)-A)*B*U))
x =
      (3*exp(t))/2 - exp(-t)/2
- (exp(-2*t)*(exp(3*t) - 7*exp(t) + 6))/2
```

As anticipated, this result matches that in Example 8.16.

8.5.3 Solution of the State Equation via State-Transition Matrix

The solution of the homogeneous state equation

$$\dot{\mathbf{x}} = \mathbf{A}\mathbf{x}, \quad \mathbf{x}(0) = \mathbf{x}_0$$

can be expressed as

$$\mathbf{x}(t) = \Phi(t)\mathbf{x}_0 \tag{8.41}$$

so that $\mathbf{x}(t)$ is obtained by a transformation of the initial state vector $\mathbf{x}_0$. For this reason, $\Phi(t)$ is known as the state-transition matrix. We next prove that $\Phi(t)$ is the unique solution of

$$\dot{\Phi} = \mathbf{A}\Phi, \quad \Phi(0) = \mathbf{I}$$

First, using Equation 8.41, we have

$$\mathbf{x}(0) = \Phi(0)\mathbf{x}_0 \quad \Rightarrow \quad \Phi(0) = \mathbf{I}$$

Next, differentiation of Equation 8.41 with respect to t , and further manipulation, yields

$$\dot{\mathbf{x}}(t) = \dot{\Phi}(t)\mathbf{x}_0 \stackrel{\Phi=\mathbf{A}\Phi}{=} \mathbf{A}\Phi(t)\mathbf{x}_0 \stackrel{\Phi(t)\mathbf{x}_0=\mathbf{x}}{=} \mathbf{A}\mathbf{x}$$

which completes the proof. Comparing Equation 8.41 with the homogeneous portions of Equations 8.40 and 8.37 reveals that

$$\Phi(t) = \mathcal{L}^{-1}\{(s\mathbf{I} - \mathbf{A})^{-1}\} = e^{\mathbf{A}t}$$

Based on this, the solution of the nonhomogeneous state equation

$$\dot{\mathbf{x}} = \mathbf{A}\mathbf{x} + \mathbf{B}\mathbf{u}, \quad \mathbf{x}(0) = \mathbf{x}_0$$

is expressed as

$$\mathbf{x}(t) = \Phi(t)\mathbf{x}_0 + \int_0^t \Phi(t-\tau)\mathbf{B}\mathbf{u}(\tau) d\tau \quad (8.42)$$

PROBLEM SET 8.4

▲ In Problems 1 through 3, find $e^{\mathbf{A}t}$, where t is scalar, using

- The `expm` command.
- The inverse Laplace-transform approach.

1. $\mathbf{A} = \begin{bmatrix} 0 & 4 \\ 1 & 3 \end{bmatrix}$

2. $\mathbf{A} = \begin{bmatrix} -2 & 1 \\ 0 & 2 \end{bmatrix}$

3. $\mathbf{A} = \begin{bmatrix} 1 & 1 & 0 \\ 1 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix}$

4. ▲ Consider the state equation subjected to an initial state vector described by

$$\dot{\mathbf{x}} = \begin{bmatrix} 3 & -1 \\ -4 & 3 \end{bmatrix} \mathbf{x} + \begin{bmatrix} 0 \\ 1 \end{bmatrix} u, \quad \mathbf{x}(0) = \begin{Bmatrix} 0 \\ -\frac{1}{2} \end{Bmatrix}$$

where:

$$u = \text{unit-pulse} = \begin{cases} \frac{1}{2} & \text{if } 0 < t < 2 \\ 0 & \text{otherwise} \end{cases}$$

Find the state vector $\mathbf{x}$ by using the formal-solution approach.

▲ In Problems 5 through 8, find the state vector via the formal-solution approach.

5. $\dot{\mathbf{x}} = \begin{bmatrix} 1 & -1 \\ -4 & 1 \end{bmatrix} \mathbf{x} + \begin{bmatrix} 1 \\ \frac{1}{2} \end{bmatrix} u, \quad u = \text{unit-step}, \quad \mathbf{x}(0) = \begin{Bmatrix} -1 \\ 0 \end{Bmatrix}$

6. $\dot{\mathbf{x}} = \begin{bmatrix} 5 & 1 \\ -4 & 1 \end{bmatrix} \mathbf{x} + \begin{bmatrix} 1 \\ -1 \end{bmatrix} u, \quad u = \sin t, \quad \mathbf{x}(0) = \begin{Bmatrix} 0 \\ 1 \end{Bmatrix}$

$$\begin{aligned}
7. \quad \dot{\mathbf{x}} &= \begin{bmatrix} 0 & 1 \\ 0 & -\frac{3}{2} \end{bmatrix} \mathbf{x} + \begin{bmatrix} 0 \\ 1 \end{bmatrix} u, \quad u = \text{unit-ramp}, \quad \mathbf{x}(0) = \begin{Bmatrix} 1 \\ 1 \end{Bmatrix} \\
8. \quad \dot{\mathbf{x}} &= \begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ -1 & -2 & -3 \end{bmatrix} \mathbf{x} + \begin{bmatrix} 0 \\ 0 \\ \frac{1}{4} \end{bmatrix} u, \quad u = e^{-t/3} \sin t, \quad \mathbf{x}(0) = \begin{Bmatrix} 0 \\ 1 \\ 0 \end{Bmatrix}
\end{aligned}$$

▲ In Problems 9 and 10, the state-space form of a system model is specified. Using the formal-solution approach, find the output $y(t)$.

$$\begin{aligned}
9. \quad \begin{cases} \dot{\mathbf{x}} = \mathbf{Ax} + \mathbf{Bu} \\ y = \mathbf{Cx} \end{cases}, \quad \mathbf{A} &= \begin{bmatrix} 0 & 1 \\ -2 & -2 \end{bmatrix}, \quad \mathbf{B} = \begin{bmatrix} 1 \\ -\frac{1}{2} \end{bmatrix}, \quad \mathbf{C} = \begin{bmatrix} 1 & 1 \end{bmatrix}, \quad u = \sin t, \quad \mathbf{x}_0 = \begin{Bmatrix} 0 \\ -1 \end{Bmatrix} \\
10. \quad \begin{cases} \dot{\mathbf{x}} = \mathbf{Ax} + \mathbf{Bu} \\ y = \mathbf{Cx} \end{cases}, \quad \mathbf{A} &= \begin{bmatrix} 2 & 0 & 2 \\ 0 & 0 & 1 \\ 0 & 0 & 4 \end{bmatrix}, \quad \mathbf{B} = \begin{bmatrix} 0 & 0 \\ 1 & 0 \\ 0 & 1 \end{bmatrix}, \quad \mathbf{C} = \begin{bmatrix} 1 & 1 & 0 \end{bmatrix}, \\
\mathbf{u} &= \begin{Bmatrix} e^{-t/2} \\ 1 \end{Bmatrix}, \quad \mathbf{x}_0 = \begin{Bmatrix} 0 \\ 0 \\ 1 \end{Bmatrix}
\end{aligned}$$

8.6 Response of Nonlinear Systems ▲

So far in this chapter, we have mainly focused on the response of linear systems. In [Section 4.6](#), we saw that the linearized model of a nonlinear system is reasonably accurate only in a small neighborhood of an operating (equilibrium) point. In order to avoid the obvious limitations that come with this approach, we tackle the nonlinear model directly, either numerically or via the simulation of the Simulink model of the system.

8.6.1 Numerical Solution of the State-Variable Equations

Mathematical models of dynamic systems of any order can be expressed in terms of state-variable equations, which may generally be nonlinear (see [Section 4.2](#)). If a system model has n state variables $x_1, x_2, \dots, x_n$ and m inputs $u_1, u_2, \dots, u_m$, the state-variable equations take the general form

$$\begin{cases} \dot{x}_1 = f_1(x_1, \dots, x_n; u_1, \dots, u_m; t) \\ \dot{x}_2 = f_2(x_1, \dots, x_n; u_1, \dots, u_m; t) \\ \dots \\ \dot{x}_n = f_n(x_1, \dots, x_n; u_1, \dots, u_m; t) \end{cases}$$

where $f_1, f_2, \dots, f_n$ are algebraic functions of the state variables and inputs. These equations may be conveniently expressed in vector form, as

$$\dot{\mathbf{x}} = \mathbf{f}(\mathbf{x}, \mathbf{u}, t) \quad (8.43)$$

where:

$$\mathbf{x} = \begin{Bmatrix} x_1 \\ x_2 \\ \dots \\ x_n \end{Bmatrix}_{n \times 1}, \quad \mathbf{u} = \begin{Bmatrix} u_1 \\ u_2 \\ \dots \\ u_m \end{Bmatrix}_{m \times 1}, \quad \mathbf{f} = \begin{Bmatrix} f_1 \\ f_2 \\ \dots \\ f_n \end{Bmatrix}_{n \times 1}$$

If the system is linear, then Equation 8.43 reduces to the familiar state equation $\dot{\mathbf{x}} = \mathbf{Ax} + \mathbf{Bu}$.

8.6.1.1 Fourth-Order Runge-Kutta Method

Among several methods designed for the numerical solution of Equation 8.43, the most commonly used is the very efficient and robust fourth-order Runge-Kutta (RK4) method. To introduce this method, we re-formulate the problem as

$$\dot{\mathbf{x}} = \mathbf{f}(t, \mathbf{x}), \quad \mathbf{x}(a) = \mathbf{x}_0, \quad a \leq t \leq b \quad (8.44)$$

with all variables as defined previously. Divide the interval $[a, b]$ into N evenly-spaced subintervals so that the mesh points are $a = t_0, t_1, t_2, \dots, t_N = b$. Let $h = (b - a)/N$ denote the step size. Knowing the initial state vector $\mathbf{x}_0$, the solution vector $\mathbf{x}_i$ at each of the subsequent mesh points t_i is obtained via

$$\mathbf{x}_{i+1} = \mathbf{x}_i + \frac{1}{6}h(\mathbf{k}_1 + 2\mathbf{k}_2 + 2\mathbf{k}_3 + \mathbf{k}_4), \quad i = 0, 1, 2, \dots, N-1 \quad (8.45)$$

where:

$$\begin{aligned} \mathbf{k}_1 &= \mathbf{f}(t_i, \mathbf{x}_i) \\ \mathbf{k}_2 &= \mathbf{f}\left(t_i + \frac{1}{2}h, \mathbf{x}_i + \frac{1}{2}h\mathbf{k}_1\right) \\ \mathbf{k}_3 &= \mathbf{f}\left(t_i + \frac{1}{2}h, \mathbf{x}_i + \frac{1}{2}h\mathbf{k}_2\right) \\ \mathbf{k}_4 &= \mathbf{f}(t_i + h, \mathbf{x}_i + h\mathbf{k}_3) \end{aligned}$$

The following user-defined function numerically solves a system of state-variable equations subjected to a prescribed initial state over a specified time interval. Note that the function is also capable of handling a scalar initial-value problem.

```
function x = RK4System(f,t,x0)
%
% RK4System uses RK4 method to solve a system of first-order
% initial-value problems in the form xdot = f(t,x), x(t0) = x0.
%
%      x = RK4System(f,t,x0), where
%
%      f is an anonymous m-dim. vector function representing f(t,x),
%      t is an (n+1)-dim. vector representing the mesh points,
%      x0 is an m-dim. vector representing the initial state vector,
%
%      x is an m-by-(n+1) matrix, each column the vector of solution
%      estimates at a mesh point.
```



```

%
x(:,1) = x0; % The first column is set to be the initial vector u0
h = t(2)-t(1); n = length(t);
for i = 1:n-1,
    k1 = f(t(i),x(:,i));
    k2 = f(t(i)+h/2,x(:,i)+h*k1/2);
    k3 = f(t(i)+h/2,x(:,i)+h*k2/2);
    k4 = f(t(i)+h,x(:,i)+h*k3);
    x(:,i+1) = x(:,i)+h*(k1+2*k2+2*k3+k4)/6;
end

```

Example 8.18: Nonlinear System Response via Fourth-Order Runge-Kutta Method

The nonlinear spring in the mechanical system shown in Figure 8.30 is identified by the spring force $f_{\text{spr}} = x^3$. The system is subjected to a unit-step externally applied force $u(t)$, and initial conditions $x(0)=0$, $\dot{x}(0)=\frac{2}{3}$.

- Find and express the state-variable equations in vector form.
- Using RK4, plot the displacement $x(t)$ versus $0 \leq t \leq 10$ sec.
- Derive the linearized model analytically about the operating point (Section 4.6), plot the response $x(t)$, and compare with the result of (b).

Solution

- The system's equation of motion is

$$\ddot{x} + \dot{x} + x^3 = u(t), \quad x(0)=0, \quad \dot{x}(0)=\frac{2}{3}$$

With state variables $x_1 = x$ and $x_2 = \dot{x}$, the nonlinear state-variable equations are obtained as

$$\begin{cases} \dot{x}_1 = x_2 \\ \dot{x}_2 = -x_2 - x_1^3 + u(t) \end{cases} \quad \begin{matrix} x_1(0)=0 \\ x_2(0)=\frac{2}{3} \end{matrix}$$

In vector form,

$$\dot{\mathbf{x}} = \mathbf{f}(t, \mathbf{x}), \quad \mathbf{x} = \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix}, \quad \mathbf{f} = \begin{Bmatrix} x_2 \\ -x_2 - x_1^3 + 1 \end{Bmatrix}, \quad \mathbf{x}_0 = \begin{Bmatrix} 0 \\ \frac{2}{3} \end{Bmatrix}, \quad 0 \leq t \leq 10$$

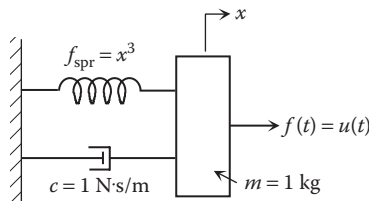


FIGURE 8.30
Nonlinear mechanical system in Example 8.18.

b. 

```

>> t = linspace(0,10); x0 = [0;2/3];
>> f = @(t,x) ([x(2);-x(2)-x(1)^3+1]);
>> x = RK4System(f, t, x0);
>> x1 = x(1,:); % Extract x1 from the state vector
>> plot(t,x1) % Initiate Figure 8.31
>> hold on % Response in (c) to be added

```

c. 

Consider the equation of motion and the initial conditions cited in (a). Following the linearization procedure outlined in Section 4.6, the operating point is obtained as $\bar{x}=1$. Linearizing the nonlinear element x^3 about the operating point yields $x^3 \cong 1+3\Delta x$, where $\Delta x = x - \bar{x} = x - 1$. The ensuing linearized model is then formulated as

$$\Delta \ddot{x} + \Delta \dot{x} + 3\Delta x = 0, \Delta x(0) = -1, \Delta \dot{x}(0) = \frac{2}{3} \quad (8.46)$$

The corresponding state-variable equations are

$$\Delta \dot{\mathbf{x}} = \mathbf{f}(t, \Delta \mathbf{x}), \Delta \mathbf{x} = \begin{Bmatrix} \Delta x_1 = \Delta x \\ \Delta x_2 = \Delta \dot{x} \end{Bmatrix}, \mathbf{f} = \begin{Bmatrix} \Delta x_2 \\ -\Delta x_2 - 3\Delta x_1 \end{Bmatrix}, \Delta \mathbf{x}_0 = \begin{Bmatrix} -1 \\ \frac{2}{3} \end{Bmatrix}, 0 \leq t \leq 10$$

Noting that $x = \Delta x + 1$, the plot of Δx_1 must be raised by 1 unit to obtain a variable compatible with x .

```

>> t = linspace(0,10); x0 = [-1;2/3];
>> f = @(t,x) ([x(2);-x(2)-3*x(1)]);
>> x = RK4System(f, t, x0);
>> x1L = x(1,:); % Extract linearized x1L from the state vector
>> plot(t,x1L+1) % Raise by 1 unit and complete Figure 8.31

```

It is immediately observed that the nonlinear and linear response curves tend to agree only in a small neighborhood of the operating point $\bar{x} = 1$.

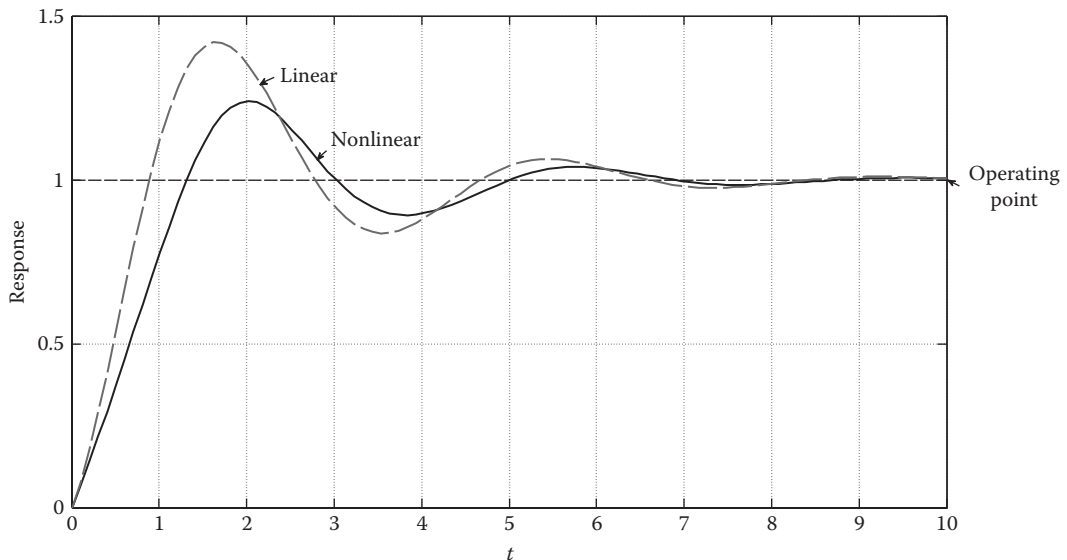


FIGURE 8.31
Responses of nonlinear and linearized models in Example 8.18.

8.6.2 Response Using Simulink Models

In [Section 4.5](#), we learned how to build a Simulink model based on the block diagram representation of the mathematical model of a dynamic system. Running the simulation of this model will generate the system response to a specified forcing function, as well as a set of prescribed initial conditions. This approach can be easily applied to nonlinear system models. The nonlinear elements are generally handled by the Fcn block in the User-Defined Functions library. In [Section 4.6](#), we learned how to linearize a nonlinear system about an (steady-state) operating point in Simulink by using Linear Analysis Points. Simulation yields the response of the linearized model. These are addressed in [Example 8.19](#) that follows.

Example 8.19: Nonlinear and Linear System Responses in Simulink

Consider the nonlinear system in [Example 8.18](#).

- Plot the response x by simulating the Simulink model of the nonlinear system.
- Linearize in Simulink and plot the response of the linearized model. Plot the nonlinear and linear response curves in a single figure and compare.

Solution

- The Simulink model shown in [Figure 8.32](#) is built based on the state-variable equations derived in [Example 8.18](#). Running the simulation and double-clicking on the Scope block reveals the response curve. As always, for direct access to the returned data, simply type

```
>> plot(tout,yout) % Initiate Figure 8.34
```

- We will linearize the model in [Figure 8.32](#) by using Linear Analysis Points. The linearization input point is placed to the left of the first integrator, the linearization output point at the output signal x_1 ; [Figure 8.33](#). The linear model is extracted from the nonlinear system as follows. Note that the model, 'Example819b', must be open before we can proceed.

```
>> sys='Example819b';
>> load_system(sys);
>> opspec = operspec(sys);

% Specify the properties of the first state variable
>> opspec.States(1).SteadyState = 1;
>> opspec.States(1).x = 0; % Initial value
>> opspec.States(1).Min = 0; % Minimum value
```

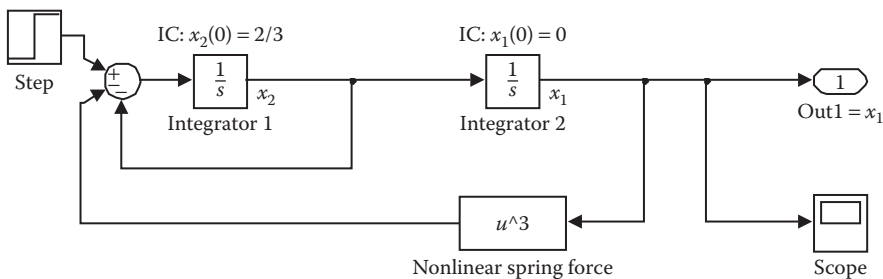


FIGURE 8.32
Simulink model in [Example 8.19\(a\)](#).


```

b =
    Integrator 1    1
    Integrator 2    0

c =
    Integrator 1    Integrator 2
    x1              0          1

d =
    x1    0

```

Continuous-time state-space model.

The state-space matrices thus obtained are of course the controller canonical form (Section 4.4) associated with the linear model, Equation 8.46, obtained in Example 8.18(c). The response of this model can be generated using the `initial` command. Note that the initial state vector is adjusted to agree with the order of the state variables demanded by the controller canonical form. Also note that the response needs to be raised by one unit to make it compatible with response x of the nonlinear system.

```

>> A = [-1 -3; 1 0]; C = [0 1];
>> sys = ss(A, [], C, []);
>> x0 = [2/3; -1];
>> [init_resp,t] = initial(sys,x0);
% Raise by 1 unit for compatibility with x and complete Figure 8.34
>> plot(t,init_resp+1)

```

As expected, Figure 8.34 agrees exactly with Figure 8.31.

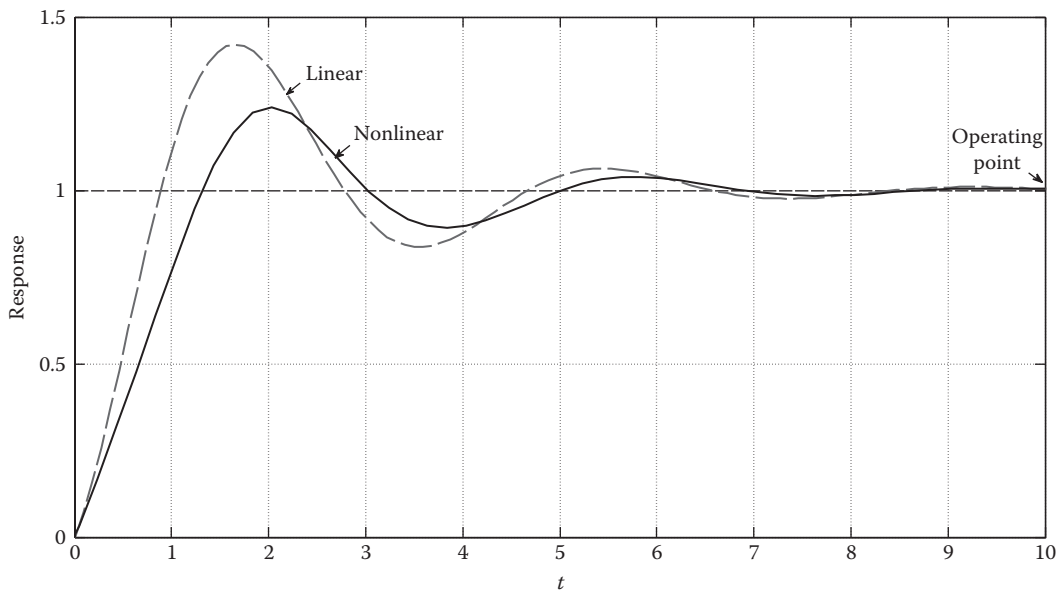


FIGURE 8.34

Linear and nonlinear responses in Example 8.19 by using Simulink.

PROBLEM SET 8.5

▲ In Problems 1 through 6, plot the specified output by using the RK4 method.

1. $3\ddot{x} + \dot{x} + x^3 = \frac{4}{3}$, $x(0) = \frac{1}{3}$, $\dot{x}(0) = 0$, $0 \leq t \leq 20$, output: $\dot{x}$
2. $2\ddot{x} + 9\dot{x} + 10x + 3x^2 = e^{-2t/5}$, $x(0) = 0$, $\dot{x}(0) = -\frac{1}{4}$, $\ddot{x}(0) = 0$, $0 \leq t \leq 15$, output: x
3. $\ddot{x} + 2\dot{x} + x\sqrt{|x|} = 1 + \sin t$, $x(0) = 1$, $\dot{x}(0) = 0$, $0 \leq t \leq 10$, output: $\dot{x}$
4. $4\ddot{x} + 3\dot{x} + 2x|x| = \frac{1}{2} + \sin t$, $x(0) = 1$, $\dot{x}(0) = -1$, $0 \leq t \leq 15$, output: x

$$5. \begin{cases} \dot{x}_1 = x_2 & x_1(0) = 1 \\ \dot{x}_2 = -\frac{1}{3}x_1 - x_1|x_1| - x_2^3 - 1 + \frac{2}{3}e^{-t/3} & x_2(0) = -\frac{1}{2} \end{cases}, \quad 0 \leq t \leq 10, \text{ output: } x_1$$

$$6. \begin{cases} \dot{x}_1 = -2x_1 - x_2^3 + e^{-t} \cos t & x_1(0) = 0 \\ \dot{x}_2 = x_1 + 4 & x_2(0) = \frac{2}{3} \end{cases}, \quad 0 \leq t \leq 5, \text{ outputs: } x_1, x_2$$

7. ▲ A nonlinear dynamic system is described by

$$\begin{cases} \dot{x}_1 = -x_2 - x_1|x_1| - 1 & x_1(0) = 0 \\ \dot{x}_2 = x_1 - x_2 - 1 & x_2(0) = -\frac{1}{3} \end{cases}, \quad 0 \leq t \leq 10$$

- a. Use the Simulink model to plot $x_2(t)$.
- b. Derive the linearized model analytically, build its Simulink model where the linear model is represented by the State-Space block, and use it to plot the time variations of the variable that is compatible with $x_2(t)$. Compare the plots generated in (a) and (b) and comment.
8. ▲ Repeat Problem 7 when x_1 is the output.
9. ▲ A nonlinear dynamic system is governed by

$$\begin{cases} \dot{x}_1 = x_2 - x_1 & x_1(0) = \frac{2}{3} \\ \dot{x}_2 = 2x_2^{-1} + 1 & x_2(0) = -\frac{1}{2} \end{cases}, \quad 0 \leq t \leq 10$$

- a. Use the Simulink model to plot $x_2(t)$.
- b. Derive the linearized model analytically, and use the `initial` command to plot the time variations of the variable that is compatible with $x_2(t)$. Compare the plots generated in (a) and (b) and comment. Note that the `initial` command will automatically determine the most suitable time duration for the response curve.
10. ▲ The mathematical model of a nonlinear system is given in the following, where $u(t)$ is the unit-step input.

$$\begin{cases} \dot{x}_1 = x_2 & x_1(0) = 0.4 \\ \dot{x}_2 = -x_1 - x_1|x_1| - x_2 + u(t) & x_2(0) = -1 \end{cases}, \quad 0 \leq t \leq 10$$

- a. Plot $x_2(t)$ by simulating the Simulink model.
- b. Derive the linearized model analytically, and use the `initial` command to plot the time variations of the variable that is compatible with $x_2(t)$. Compare the plots generated in (a) and (b) and comment.

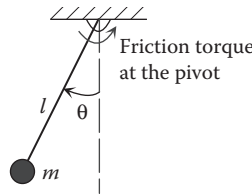


FIGURE 8.35
Problem 12.

- c. Find the linear model by using the linearization procedure in Simulink. Plot the response $x_2(t)$ to confirm the result of (b).

11. Consider the nonlinear model

$$\dot{x} + ax|x| = e^{-2t/3} \sin t, \quad x(0) = 0.3, \quad 0 \leq t \leq 15$$

where a is a parameter. Use the RK4 method to plot the response $x(t)$ for $a = 0.2, 0.8, 1.5$ versus t in the same graph.

12. The pendulum system in Figure 8.35 consists of a uniform thin rod of length l and a concentrated mass m at its tip. The friction at the pivot causes the system to be damped. When the angular displacement θ is not very small, the system is described by the nonlinear model

$$ml^2\ddot{\theta} + 0.24\dot{\theta} + \frac{1}{2}mgl \sin \theta = 0$$

Assume, in consistent physical units, that $ml^2 = 1.5$, $\frac{g}{l} = 6.8$. Two sets of initial conditions are to be considered: (1) $\theta(0) = 10^\circ$, $\dot{\theta}(0) = 0$, and (2) $\theta(0) = 20^\circ$, $\dot{\theta}(0) = 0$. Using the RK4 method plot the two angular displacements θ_1 and θ_2 corresponding to the two sets of initial conditions versus $0 \leq t \leq 10$ in the same graph. Angle measures must be converted to radians.

13. A nonlinear dynamic system model is derived as

$$\dot{x} + 2x^3 = u(t), \quad x(0) = 0, \quad 0 \leq t \leq 3$$

where $u(t)$ is the unit-step function.

- Build the Simulink model and use it to plot the response $x(t)$.
 - Derive the linearized model analytically. Build a Simulink model and use it to plot the time variations of the variable in the linear model that is compatible with $x(t)$. Compare the plots generated in (a) and (b) and comment.
14. The nonlinear spring in the mechanical system shown in Figure 8.36 is identified by the spring force $f_{\text{spr}} = x|x|$. The system is subjected to a unit-step externally applied force $u(t)$, and initial conditions $x(0) = 0$, $\dot{x}(0) = 1$.
- Derive the state-variable equations in vector form.
 - Using the RK4 method, plot the displacement $x(t)$ versus $0 \leq t \leq 10$ sec.

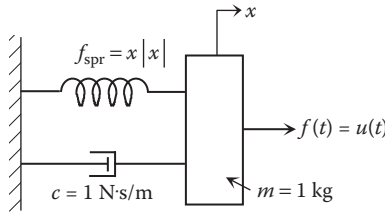


FIGURE 8.36
Problem 14.

8.7 Summary

The total response of a dynamic system consists of transient and steady-state responses. The transient response comprises those terms in the total response that decay to zero after a sufficiently long time. The portion of the total response that remains after the transient terms have vanished describes the steady-state response. Linear, first-order systems are modeled as

$$\tau \dot{x} + x = f(t), \quad \tau = \text{time const} > 0, \quad x(0) = x_0$$

where the time constant provides a measure of how quickly the response reaches a steady state. Linear, second-order systems are modeled as

$$\ddot{x} + 2\zeta\omega_n\dot{x} + \omega_n^2x = f(t), \quad x(0) = x_0, \quad \dot{x}(0) = \dot{x}_0$$

where ζ and ω_n are the damping ratio and (undamped) natural frequency, respectively.

The MATLAB function `initial` finds system response to initial conditions, while `impz` and `step` built-in functions find the responses to impulse and step inputs, assuming zero initial conditions. For cases of general input and/or non-zero initial conditions, the `lsim` command is used. Response analysis may also be performed through simulation of the Simulink model of the system at hand.

A system's steady-state response to a sinusoidal input is called the frequency response. The frequency response of a system with transfer function $G(s)$ to input $F_0 \sin \omega t$ is

$$F_0 |G(j\omega)| \sin(\omega t + \phi), \quad \phi = \angle G(j\omega)$$

where $G(j\omega)$ is the FRF.

A Bode plot consists of a pair of plots: $20 \log |G(j\omega)|$ with the unit of decibel (dB) versus frequency, and the phase ϕ in degrees versus frequency. The curves are plotted on semilog paper, using a linear scale for magnitude (in dB) and phase (in degrees) and a logarithmic scale for frequency (in rad/sec). The formal solution of the state equation

$$\dot{\mathbf{x}} = \mathbf{Ax} + \mathbf{Bu}, \quad \mathbf{x}(0) = \mathbf{x}_0$$

is expressed as

$$\mathbf{x}(t) = e^{\mathbf{A}t} \mathbf{x}_0 + \int_0^t e^{\mathbf{A}(t-\tau)} \mathbf{B} \mathbf{u}(\tau) d\tau$$

where $e^{\mathbf{A}t}$ is the matrix exponential of $\mathbf{A}$. The MATLAB built-in function `expm` handles the matrix exponential.

The RK4 method numerically solves a set of state-variable equations

$$\dot{\mathbf{x}} = \mathbf{f}(t, \mathbf{x}), \quad \mathbf{x}(a) = \mathbf{x}_0, \quad a \leq t \leq b$$

where:

$\mathbf{x}_0$ is the initial state vector

$[a, b]$ is the interval in which the system is solved

For an integer $N > 0$ and step size $h = (b - a)/N$, the mesh points $t_i = a + ih, i = 0, 1, 2, \dots, N - 1$, partition $[a, b]$ into N subintervals. Starting with $\mathbf{x}_0$, the solution vector $\mathbf{x}_i$ at each of the subsequent mesh points t_i is obtained via

$$\mathbf{x}_{i+1} = \mathbf{x}_i + \frac{1}{6} h (\mathbf{k}_1 + 2\mathbf{k}_2 + 2\mathbf{k}_3 + \mathbf{k}_4), \quad i = 0, 1, 2, \dots, N - 1$$

where:

$$\mathbf{k}_1 = \mathbf{f}(t_i, \mathbf{x}_i)$$

$$\mathbf{k}_2 = \mathbf{f}(t_i + \frac{1}{2}h, \mathbf{x}_i + \frac{1}{2}h\mathbf{k}_1)$$

$$\mathbf{k}_3 = \mathbf{f}(t_i + \frac{1}{2}h, \mathbf{x}_i + \frac{1}{2}h\mathbf{k}_2)$$

$$\mathbf{k}_4 = \mathbf{f}(t_i + h, \mathbf{x}_i + h\mathbf{k}_3)$$

Review Problems

1. A linear system is modeled as

$$\frac{2}{5} \dot{x} + 3x = \frac{5}{3} u(t) + u_r(t), \quad x(0) = 0$$

where $u(t)$ and $u_r(t)$ denote the unit-step and unit-ramp functions, respectively.

- a. Express the response $x(t)$ in closed form.

- b.  Plot the response $x(t)$ by using the `lsim` command.

2. The model of a second-order system is given as

$$4\ddot{x} + 4\dot{x} + \frac{5}{4}x = 0, \quad x(0) = 0.5, \quad \dot{x}(0) = -1$$

- a. Identify the damping type and express the response $x(t)$ in closed form.
 - b.  Plot the response $x(t)$ by using the `initial` command.
3. A second-order dynamic system is modeled as

$$9\ddot{x} + 6\dot{x} + \frac{10}{9}x = 14\delta(t), \quad x(0) = 0, \quad \dot{x}(0) = -\frac{1}{4}$$

- a. Find the response $x(t)$ in closed form.
 - b.  Plot the response $x(t)$ by using the `impz` and `initial` commands. Also, plot the impulse and initial responses in the same figure.
4.  The mathematical model of a dynamic system is described by

$$4\ddot{x} + 4\dot{x} + 5x = \frac{10}{3}u_r(t), \quad x(0) = 0, \quad \dot{x}(0) = \frac{2}{3}$$

where $u_r(t)$ is the unit ramp. Plot the response $x(t)$ by

- a. Using the `lsim` command over $0 \leq t \leq 5$.
 - b. Simulating the Simulink model of the system.
5.  A dynamic system model is given as

$$\begin{cases} 2\ddot{x}_1 + \dot{x}_1 - \frac{1}{3}(x_2 - x_1) = u(t) \\ \dot{x}_2 + x_2 + \frac{1}{3}(x_2 - x_1) = \frac{5}{2}u(t) \end{cases}, \quad x_1(0) = 1, \quad x_2(0) = 0, \quad \dot{x}_1(0) = \frac{1}{5}$$

where $u(t)$ is the unit step. Plot x_1 versus $0 \leq t \leq 15$ by

- a. Using the `lsim` command.
 - b. Simulating the Simulink model of the system.
6.  The governing equations for a system are derived as

$$\begin{cases} 3\ddot{x}_1 + \dot{x}_1 + 3x_1 - 2x_2 = 10\delta(t) \\ \dot{x}_2 - 3x_1 + 3x_2 = 0 \end{cases}, \quad x_1(0) = 1, \quad x_2(0) = 0, \quad \dot{x}_1(0) = 0$$

where $\delta(t)$ is the unit impulse. Plot x_1 by using the `impz` and `initial` commands.

7.  The mechanical system in [Figure 8.37](#), where all parameter values are in consistent physical units, is subject to initial conditions $x_1(0) = 1$, $x_2(0) = 1$, $\dot{x}_1(0) = -1$, $\dot{x}_2(0) = 1$. Plot the response x_1 versus $0 \leq t \leq 25$.

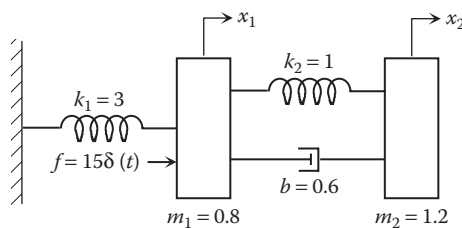


FIGURE 8.37
Problem 7.

8. The Bode plot of a first-order system indicates that the corner frequency is 0.35 rad/sec and high-frequency (at $\omega = 100$ rad/sec) logarithmic magnitude is -37.45 dB.
 - a. Find the system's transfer function.
 - b. Using the results of (a), find the approximate low-frequency logarithmic magnitude.
9. Determine the damping ratio associated with a second-order system in the standard form of Equation 8.32 that corresponds to a maximum (peak) logarithmic magnitude of 15.22 dB.
10. The Bode plot of a second-order system indicates that the approximate low-frequency magnitude is 2.28 dB, the approximate high-frequency ($\omega = 1000$ r/s) magnitude is -106 dB, and the logarithmic gain attains a peak value of 4.98 dB. Find the system's transfer function.
11.  Consider

$$\begin{cases} \dot{x}_1 = -x_1|x_1| + ax_2 - \frac{2}{3} + \sin t & x_1(0) = \frac{2}{3} \\ \dot{x}_2 = -x_1 - \frac{1}{2}x_2 - 1 & x_2(0) = 1 \end{cases}$$

where a is a parameter. Using the RK4 method, plot x_1 versus $0 \leq t \leq 10$ for the two cases $a = 1.25$ and $a = 2.5$ in the same graph.

12.  The nonlinear state-variable equations for a dynamic system are derived as

$$\begin{cases} \dot{x}_1 = -2x_2 + 3.3\sin t & x_1(0) = 0.7 \\ \dot{x}_2 = x_1^3 - 3x_2 + 5 & x_2(0) = 0 \end{cases}$$

Plot $x_1(t)$ versus $0 \leq t \leq 10$ by

- a. Using the RK4 method.
- b. Simulating the Simulink model of the system.

9

Introduction to Vibrations

Vibration can be regarded as a subset of dynamics, in which a system subjected to restoring forces oscillates about an equilibrium position. The restoring forces are due to elasticity or gravity. Two different types of excitations cause a system to vibrate: initial excitation and external excitation. The vibration of a system caused by non-zero initial excitations, including initial displacements or initial velocities (or both), is known as free vibration. The vibration of a system caused by externally applied forces is known as forced vibration.

In this chapter, we first extend the knowledge gathered in [Chapters 5](#) and [8](#) to the analysis of free vibration and forced vibration of single-degree-of-freedom or two-degree-of-freedom systems. [Section 9.1](#) introduces the free vibration of Coulomb damped systems, in which energy is dissipated via dry friction. [Section 9.2](#) considers the forced vibration of systems with rotating eccentric masses and systems with harmonically moving supports. [Sections 9.1](#) and [9.2](#) also cover topics such as logarithmic decrement and bandwidth, both of which can be used to estimate the widely-used viscous damping model. To reduce the effects of undesired vibration, [Section 9.3](#) discusses the design of vibration suppression systems, including vibration isolators and vibration absorbers. For multi-degree-of-freedom systems, it is convenient to use the matrix-based method to perform vibration analysis. In [Section 9.4](#), the concepts of the eigenvalue problem, natural modes, and orthogonality of modes are presented by means of matrix algebra. The modal analysis method is developed and used to obtain the response to initial or harmonic excitations. The chapter concludes with coverage of vibration measurement technology. [Section 9.5](#) introduces the hardware available for vibration testing and the methods used to identify system parameters, such as natural frequencies and damping ratios.

9.1 Free Vibration

The governing differential equation for a single-degree-of-freedom viscously damped system subjected to non-zero initial excitations is

$$m\ddot{x}(t) + b\dot{x}(t) + kx(t) = 0 \quad (9.1)$$

or

$$\ddot{x}(t) + 2\zeta\omega_n\dot{x}(t) + \omega_n^2x(t) = 0 \quad (9.2)$$

with $x(0) = x_0$ and $\dot{x}(0) = v_0$. As indicated in [Section 8.3](#), the nature of the system response to initial excitations depends on the value of the damping ratio ζ . For an overdamped system ($\zeta > 1$), the response represents aperiodic decay, which is also true for a critically damped

system ($\zeta = 1$). For an underdamped system ($0 < \zeta < 1$), the response represents oscillatory decay. For an undamped system ($\zeta = 0$), the response represents harmonic oscillation with natural frequency ω_n .

Viscous damping is a widely used damping model, but not the only one. Damping is a very complex phenomenon and a wide range of damping models can be found in the literature. In this section, we first discuss the measurement of the viscous damping ratio ζ based on transient time response plots and then introduce another damping model, known as Coulomb damping.

9.1.1 Logarithmic Decrement

Unlike mass m and spring stiffness k , which could both be easily measured with static tests, the viscous damping coefficient b has to be measured with a dynamic test. A common way is to use the free response of the whole system to measure the damping ratio ζ and then determine the damping coefficient b by using $b = 2\zeta\sqrt{mk}$.

As discussed in [Section 8.3](#), the free response of an underdamped single-degree-of-freedom system is

$$x(t) = e^{-\zeta\omega_n t} \left(x_0 \cos \omega_d t + \frac{\zeta\omega_n x_0 + v_0}{\omega_d} \sin \omega_d t \right) \quad (9.3)$$

where the damped natural frequency is $\omega_d = \omega_n \sqrt{1 - \zeta^2}$. Equation 9.3 can also be written as (see [Section 2.2](#))

$$x(t) = Ae^{-\zeta\omega_n t} \cos(\omega_d t - \phi) \quad (9.4)$$

where the amplitude A and the phase ϕ are given by

$$A = \sqrt{x_0^2 + \left(\frac{\zeta\omega_n x_0 + v_0}{\omega_d} \right)^2} \quad (9.5)$$

and

$$\phi = \tan^{-1} \frac{\zeta\omega_n x_0 + v_0}{\omega_d x_0} \quad (9.6)$$

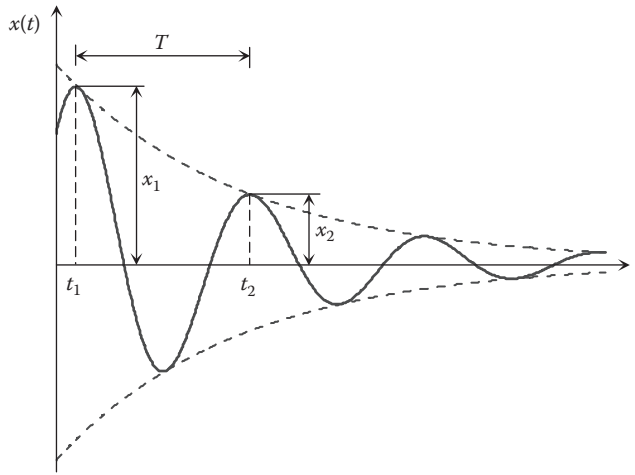
[Figure 9.1](#) shows the system response to initial excitations, where $T = 2\pi/\omega_d$ is the period of damped oscillation.

Note that the peak drops after one cycle of vibration, and the ratio between the first and the second peaks is

$$\frac{x_1}{x_2} = \frac{x(t_1)}{x(t_2)} = \frac{Ae^{-\zeta\omega_n t_1} \cos(\omega_d t_1 - \phi)}{Ae^{-\zeta\omega_n t_2} \cos(\omega_d t_2 - \phi)} = \frac{e^{-\zeta\omega_n t_1} \cos(\omega_d t_1 - \phi)}{e^{-\zeta\omega_n (t_1 + T)} \cos[\omega_d (t_1 + T) - \phi]} \quad (9.7)$$

Because $\omega_d T = 2\pi$, we have

$$\zeta\omega_n T = \zeta\omega_n \left(\frac{2\pi}{\omega_d} \right) = 2\pi\zeta/\sqrt{1 - \zeta^2} \quad \text{and} \quad \cos[\omega_d (t_1 + T) - \phi] = \cos(\omega_d t_1 + 2\pi - \phi) = \cos(\omega_d t_1 - \phi)$$

**FIGURE 9.1**

Free response of an underdamped single-degree-of-freedom system to initial excitations.

Thus, Equation 9.7 reduces to

$$\frac{x_1}{x_2} = e^{2\pi\zeta/\sqrt{1-\zeta^2}} \quad (9.8)$$

Taking the natural logarithm of both sides of Equation 9.8 yields

$$\delta = \ln \frac{x_1}{x_2} = \frac{2\pi\zeta}{\sqrt{1-\zeta^2}} \quad (9.9)$$

where δ is called the logarithmic decrement. It turns out that the ratio of any two consecutive displacement peaks gives the same result as Equation 9.8,

$$\frac{x_1}{x_2} = \frac{x_2}{x_3} = \dots = \frac{x_n}{x_{n+1}} = e^{2\pi\zeta/\sqrt{1-\zeta^2}} \quad (9.10)$$

Thus, the logarithmic decrement δ can be determined by measuring any two consecutive displacement peaks. The damping ratio ζ can be obtained from Equation 9.9, as

$$\zeta = \frac{\delta}{\sqrt{(2\pi)^2 + \delta^2}} \quad (9.11)$$

For a more accurate estimation, the damping ratio ζ can be determined by measuring the displacements of two peaks separated by a number of periods, as opposed to two consecutive peaks. If we denote the peak displacements at times t_1 and t_{n+1} as x_1 and x_{n+1} , respectively, where $t_{n+1} = t_1 + nT$, then the ratio between the two peak displacements can be written as

$$\frac{x_1}{x_{n+1}} = \frac{x_1}{x_2} \frac{x_2}{x_3} \dots \frac{x_n}{x_{n+1}} \quad (9.12)$$

Taking the natural logarithm of both sides of Equation 9.12, we have

$$\ln \frac{x_1}{x_{n+1}} = \ln \left(\frac{x_1}{x_2} \frac{x_2}{x_3} \cdots \frac{x_n}{x_{n+1}} \right) = \ln \frac{x_1}{x_2} + \ln \frac{x_2}{x_3} + \cdots + \ln \frac{x_n}{x_{n+1}} = n\delta \quad (9.13)$$

which yields the logarithmic decrement as

$$\delta = \frac{1}{n} \ln \frac{x_1}{x_{n+1}} \quad (9.14)$$

Substituting Equation 9.14 into Equation 9.11 gives the damping ratio. If the damping is very small, that is, $\zeta \ll 1$ and $\sqrt{1-\zeta^2} \approx 1$, then Equation 9.9 gives

$$\zeta \approx \frac{\delta}{2\pi} \quad (9.15)$$

Example 9.1: Logarithmic Decrement

A vibrating system consisting of a mass of 5 kg and a spring of stiffness 5400 N/m is viscously damped. The ratio of any two consecutive amplitudes is $\frac{3}{2}$.

- Determine the logarithmic decrement δ .
- Determine the exact value of the damping ratio ζ .
- Determine the damping coefficient b .
- Assuming small damping, recalculate the damping ratio ζ and determine the percentage of error.

Solution

- The logarithmic decrement can be determined by measuring the displacements of any two consecutive peaks, as

$$\delta = \ln \frac{x_j}{x_{j+1}} = \ln \frac{3}{2} = 0.4055$$

- The exact value of the damping ratio is

$$\zeta = \frac{\delta}{\sqrt{(2\pi)^2 + \delta^2}} = \frac{0.4055}{\sqrt{(2\pi)^2 + (0.4055)^2}} = 0.0644$$

- The viscous damping coefficient is

$$b = 2\zeta\sqrt{mk} = 2(0.0644)\sqrt{5(5400)} = 21.1640 \text{ N} \cdot \text{s/m}$$

- For small damping,

$$\zeta \approx \frac{\delta}{2\pi} = \frac{0.4055}{2\pi} = 0.0645$$

and the percentage of error is 0.16%. Note that the error is close to zero, and thus, the assumption of small damping is valid.

9.1.2 Coulomb Damping

The linear viscous damping model was adopted in the previous chapters. The viscous damping force is linearly dependent on velocity, simplifying the analysis. Although this is widely used in vibration, it is not the only damping model. Coulomb damping is another type of damping in which energy is dissipated via dry friction.

Figure 9.2 shows a mass–spring system subject to Coulomb damping, in which N is the normal force and F_f is the dry friction force. Let μ_k denote the kinetic friction coefficient. Note that the friction force remains constant in magnitude, which is $\mu_k mg$, and the force is opposite in direction to the motion or the velocity $\dot{x}$. Introducing the sign function $\text{sgn}(\cdot)$,

$$\text{sgn}(\dot{x}) = \begin{cases} 1, & \dot{x} > 0 \\ -1, & \dot{x} < 0 \end{cases} \quad (9.16)$$

the friction force can be expressed as

$$F_f = -F_f \text{sgn}(\dot{x}) = \begin{cases} -F_f, & \dot{x} > 0 \\ F_f, & \dot{x} < 0 \end{cases} \quad (9.17)$$

where $F_f = \mu_k mg$. Equation 9.17 implies that the friction force points to the left if the mass moves to the right and points to the right if the mass moves to the left.

Using Equation 9.17, we can write the dynamics equation of motion as

$$m\ddot{x} + F_f \text{sgn}(\dot{x}) + kx = 0 \quad (9.18)$$

which is a nonlinear equation that can be separated into two linear equations,

$$m\ddot{x} + F_f + kx = 0, \quad \dot{x} > 0 \quad (9.19)$$

$$m\ddot{x} - F_f + kx = 0, \quad \dot{x} < 0 \quad (9.20)$$

Without loss of generality, let us assume the initial conditions to be $x(0) = x_0 > 0$ and $\dot{x}(0) = v_0 = 0$. Due to the restoring spring force, the mass first moves from right to left, as shown in Figure 9.3a. The velocity is negative and the dynamics of the system is expressed by Equation 9.20, which can be rewritten as

$$m\ddot{x} + k(x - \Delta) = 0 \quad (9.21)$$

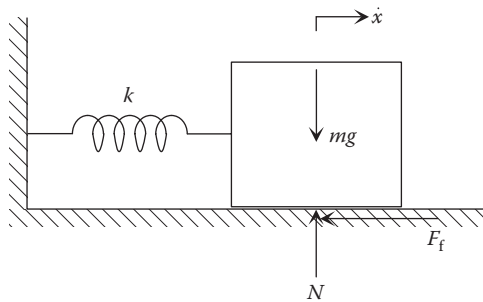
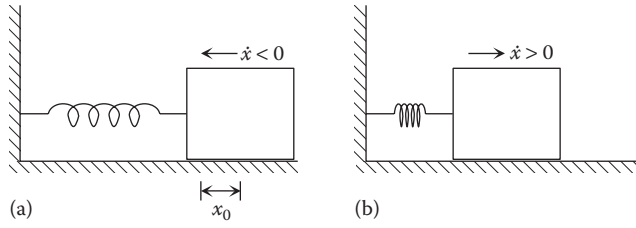


FIGURE 9.2

A mass–spring system subject to Coulomb damping.

**FIGURE 9.3**

Motion of the mass with (a) negative velocity and (b) positive velocity.

where $\Delta = F_f/k$. For a given system, the mass m , spring stiffness k , and kinetic friction coefficient μ_k are all constants. Thus, Δ is a constant and $\dot{\Delta} = 0$. Equation 9.21 can then be rewritten as

$$m(\ddot{x} - \ddot{\Delta}) + k(x - \Delta) = 0 \quad (9.22)$$

Note that Equation 9.22 has the same format as $m\ddot{x} + kx = 0$ if $x - \Delta$ is replaced with x . Therefore, the free response of the system can be determined by using Equations 9.4 through 9.6 and neglecting damping. If we let $x' = x - \Delta$, the initial conditions can be expressed as $x'(0) = x(0) - \Delta = x_0 - \Delta$ and $\dot{x}'(0) = \dot{x}(0) - \dot{\Delta} = 0$. Substituting the initial conditions into Equations 9.4 through 9.6 yields

$$x(t) - \Delta = (x_0 - \Delta) \cos(\omega_n t) \quad (9.23)$$

or

$$x(t) = (x_0 - \Delta) \cos(\omega_n t) + \Delta \quad (9.24)$$

When the spring reaches maximum compression, as shown in [Figure 9.3b](#), the velocity of the mass reduces to zero, that is, $\dot{x}(t) = -(x_0 - \Delta)\omega_n \sin(\omega_n t) = 0$, which yields $t = \pi/\omega_n = T/2$. The corresponding displacement is $-(x_0 - 2\Delta)$. The mass then starts to move from left to right and the velocity becomes positive. Thus, Equation 9.24 is only valid for $0 \leq t < T/2$.

For $t \geq T/2$, the dynamics of the system is expressed by Equation 9.19, which can be rewritten as

$$m\ddot{x} + k(x + \Delta) = 0 \quad (9.25)$$

or

$$m(\ddot{x} + \ddot{\Delta}) + k(x + \Delta) = 0 \quad (9.26)$$

The response of the system can also be determined by using Equations 9.4 through 9.6 and neglecting damping. Let $x' = x + \Delta$. Note that the initial values of x and $\dot{x}$ in Equation 9.26 are the displacement and the velocity at time $T/2$, respectively, that is, $x(T/2) = -(x_0 - 2\Delta)$ and $\dot{x}(T/2) = 0$. As a result, $x'(0) = -(x_0 - 3\Delta)$ and $\dot{x}'(0) = 0$. The solution of Equation 9.26 is

$$x(t) + \Delta = (x_0 - 3\Delta) \cos(\omega_n t) \quad (9.27)$$

or

$$x(t) = (x_0 - 3\Delta) \cos(\omega_n t) - \Delta \quad (9.28)$$

When the spring reaches maximum elongation, the velocity of the mass reduces to zero once again. By differentiating Equation 9.28 and equating it to zero, we obtain the time corresponding to the maximum elongation, that is, T . The displacement at $t = T$ is $x_0 - 4\Delta$. After this point, the motion reverses in direction and the mass moves from right to left. Thus, Equation 9.28 is only valid for $T/2 \leq t < T$.

The previous discussion gives the displacement magnitudes at $t = 0$, $T/2$, and T , which are x_0 , $x_0 - 2\Delta$, and $x_0 - 4\Delta$, respectively. It can be concluded that the displacement magnitude is reduced by 2Δ after every half-cycle. This process is repeated as the mass oscillates back and forth about its equilibrium position. The motion stops when the displacement is not large enough for the restoring spring force to overcome the static friction force. The free response of the mass–spring system with Coulomb damping is shown in Figure 9.4. The envelope of the response for a vibrating system with Coulomb damping is a straight line instead of an exponential decay curve for viscously damped systems, as shown in Figure 9.1.

Example 9.2: Coulomb Damping

For a Coulomb damped system, it is observed that the first three consecutive maximum displacement amplitudes x_0 , x_1 , and x_2 for a free vibration are 15, 12.55, and 10.10 cm, respectively. The time duration between any two of these amplitudes is 0.6 s.

- Determine the value of the kinetic friction coefficient μ_k .
- Determine the position when the oscillation stops.

Solution

- The decay per cycle is

$$4\Delta = 15 - 12.55 = 12.55 - 10.10 = 2.45 \text{ cm}$$

which gives $\Delta = 6.125 \times 10^{-3} \text{ m}$. The time duration between any two of these amplitudes is 0.6 s. This implies the period of vibration is $T = 0.6 \text{ s}$ and

$$\omega_n = \sqrt{\frac{k}{m}} = \frac{2\pi}{T} = \frac{2\pi}{0.6} = 10.47 \text{ rad/s}$$

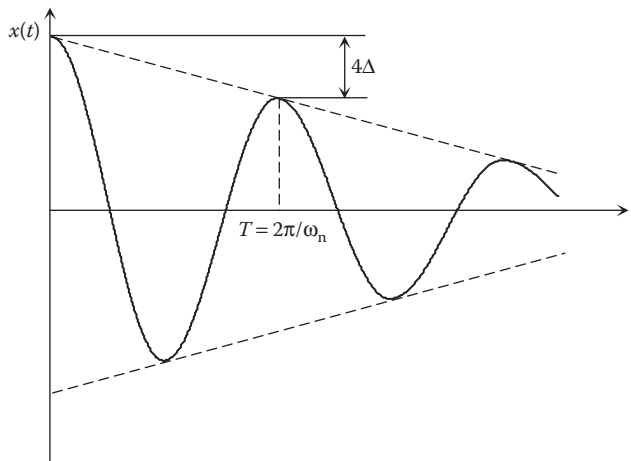


FIGURE 9.4

Free response of a mass–spring system subject to Coulomb damping.

Thus, the kinetic friction coefficient is

$$\mu_k = \frac{F_f}{mg} = \frac{k\Delta}{mg} = \frac{m\omega_n^2\Delta}{mg} = \frac{10.47^2(6.125 \times 10^{-3})}{9.81} = 0.07$$

- b. Note that the displacement magnitude is reduced by 2Δ after every half-cycle. The motion stops at the end of the half-cycle for which the displacement magnitude is smaller than 2Δ . This can be expressed mathematically as

$$x_0 - n(2\Delta) < 2\Delta$$

where n denotes the number of half-cycles before stopping. Solving for n and substituting the appropriate values, we find

$$n > \frac{x_0}{2\Delta} - 1 = \frac{0.15}{2(6.125 \times 10^{-3})} - 1 = 11.24$$

The smallest integer satisfying the inequality is $n = 12$. Thus, the oscillation stops after 12 half-cycles, or 6 cycles, with

$$x(t = 12T/2) = x_0 - 12(2\Delta) = 0.15 - 12(2 \times 6.125 \times 10^{-3}) = 3 \times 10^{-3} \text{ m}$$

PROBLEM SET 9.1

1. A lightly damped single-degree-of-freedom system is subjected to free vibration. The response of the system is shown in Figure 9.5. Estimate the value of the viscous damping ratio ζ .
2. An underdamped single-degree-of-freedom vibrating system is viscously damped. It is observed that the maximum displacement amplitude during the third cycle is

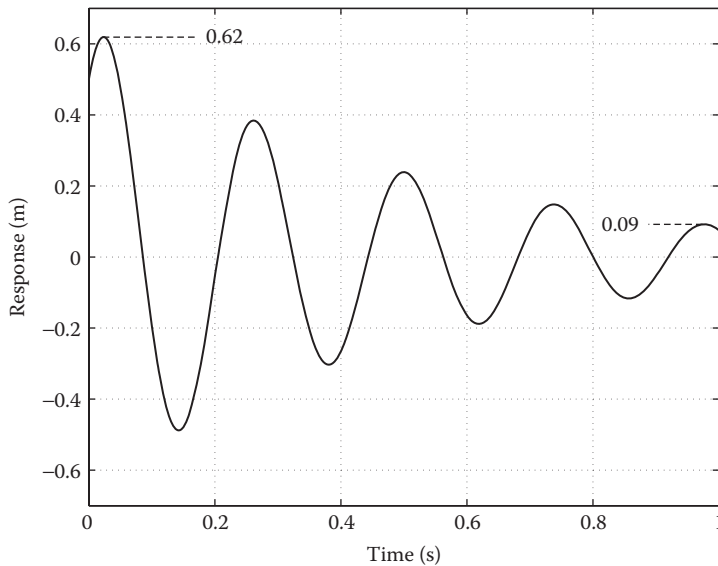


FIGURE 9.5
Problem 1.

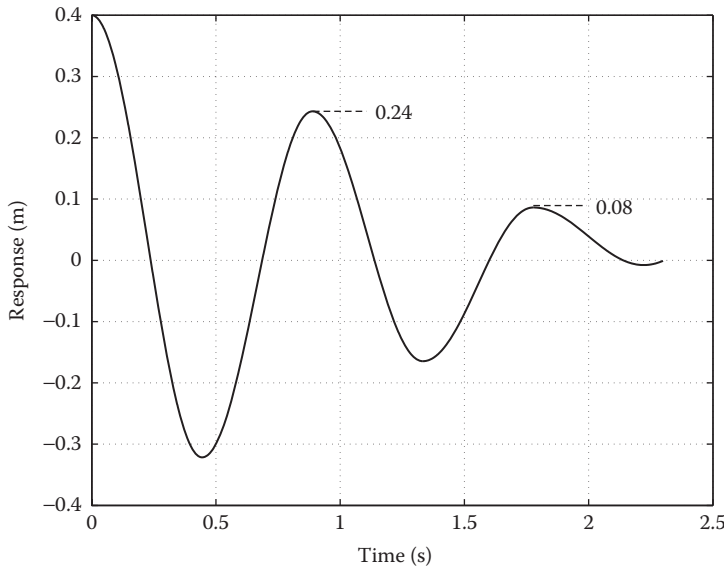


FIGURE 9.6
Problem 3.

60% of the first. Calculate the damping ratio ζ and determine the maximum displacement amplitude during the fourth cycle as a fraction of the first.

3. Figure 9.6 is the free response of a single-degree-of-freedom system subjected to Coulomb damping. The parameters of the system include the mass ($m = 40$ kg) and the spring stiffness ($k = 2000$ N/m). Estimate the value of the kinetic friction coefficient μ_k .
4. A Coulomb damped vibrating system consists of a mass of 8 kg and a spring of stiffness 5000 N/m. The kinetic friction coefficient μ_k is 0.1. The initial conditions are $x_0 = 0.02$ m and $v_0 = 0$ m/s.
 - a. Determine the decay per cycle.
 - b. Determine the position when the oscillation stops.

9.2 Forced Vibration

The vibration of a system caused by externally applied forces is known as forced vibration. A very important class of external excitations involves harmonic forces. Consider a single-degree-of-freedom viscously damped system subjected to a harmonic excitation, $f(t) = F_0 \sin(\omega t)$. The governing differential equation is given by

$$m\ddot{x}(t) + b\dot{x}(t) + kx(t) = F_0 \sin(\omega t) \quad (9.29)$$

or

$$\ddot{x}(t) + 2\zeta\omega_n\dot{x}(t) + \omega_n^2 x(t) = \omega_n^2 \frac{F_0}{k} \sin(\omega t) \quad (9.30)$$

As discussed in [Section 8.4](#), the steady-state response of a system subjected to a harmonic excitation $f(t)$ may be obtained using the frequency response, which is a very important concept in vibration. The steady-state response of the system described by Equation 9.29 or 9.30 is

$$x(t) = X \sin(\omega t + \phi) \quad (9.31)$$

where the amplitude and the phase angle are given by

$$X = \frac{F_0/k}{\sqrt{[1 - (\omega/\omega_n)^2]^2 + [2\zeta(\omega/\omega_n)]^2}} \quad (9.32)$$

and

$$\phi = -\tan^{-1} \frac{2\zeta(\omega/\omega_n)}{1 - (\omega/\omega_n)^2} \quad (9.33)$$

Note that the value of X depends on the driving frequency ω , and X is called the dynamic amplitude. The term F_0/k in Equation 9.32 has units of displacement and is known as the static deflection. If we use x_{st} to denote the static deflection, we can obtain the dimensionless ratio

$$\frac{X}{x_{st}} = \frac{1}{\sqrt{[1 - (\omega/\omega_n)^2]^2 + [2\zeta(\omega/\omega_n)]^2}} \quad (9.34)$$

When compared with other excitations, more information on the steady-state response to harmonic excitations can be extracted using the frequency domain technique rather than the time domain technique. In this section, we first discuss how to measure the viscous damping coefficient ζ based on frequency response plots. Systems subjected to two types of harmonic excitations, including rotating unbalance and base excitation, are then introduced and their responses are determined using the frequency response method.

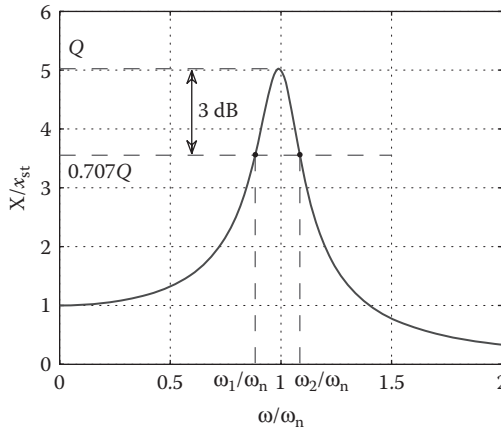
9.2.1 Half-Power Bandwidth

[Figure 9.7](#) shows the frequency response of the dimensionless ratio X/x_{st} near the natural frequency for a viscously damped system. As proven in [Section 8.4](#), the maximum peak occurs at $\omega/\omega_n = \sqrt{1 - 2\zeta^2}$. If the system is lightly damped, the peak occurs in the immediate neighborhood of $\omega/\omega_n = 1$, as shown in [Figure 9.7](#). From Equation 9.34, the corresponding value of the peak amplitude can be approximated as

$$Q = \left(\frac{X}{x_{st}} \right)_{\max} \approx \frac{1}{2\zeta} \quad (9.35)$$

for small damping. The symbol Q introduced in Equation 9.35 is known as the quality factor or Q factor, which is usually used in electrical engineering applications and is related to the amplitude at resonance. Thus, the damping ratio ζ can be obtained from Equation 9.35, as

$$\zeta \approx \frac{1}{2Q} \quad (9.36)$$

**FIGURE 9.7**

Magnitude of the frequency response for a viscously damped system.

which can be used as a quick way of estimating the viscous damping ratio by measuring the peak amplitude Q .

For a more accurate estimation, the damping ratio ζ can be determined by measuring the frequencies at two half-power points instead of the peak amplitude. Note that for a viscously damped system subjected to a harmonic force, the velocity response is

$$\dot{x}(t) = \omega X \cos(\omega t + \phi) \quad (9.37)$$

which is obtained by taking the time derivative of Equation 9.31. The maximum kinetic energy is

$$T_{\max} = \frac{1}{2} m \omega^2 X_{\max}^2 = \frac{1}{2} m \omega^2 x_{\text{st}}^2 Q^2 \quad (9.38)$$

which is proportional to the square of Q . Half-power points occur at frequencies where the power drops to half of its maximum level. As shown in Figure 9.7, points 1 and 2, at which the amplitude falls to $Q/\sqrt{2} \approx 0.707Q$, or drops by 3 dB down from the peak, are half-power points.

Let $v = \omega/\omega_n$. At the half-power points, combining Equations 9.34 and 9.35 yields

$$\frac{1}{\sqrt{(1-v^2)^2 + (2\zeta v)^2}} \approx \frac{1}{\sqrt{2}} \frac{1}{2\zeta} \quad (9.39)$$

or

$$v^4 - 2(1-2\zeta^2)v^2 + (1-8\zeta^2) = 0 \quad (9.40)$$

which has the solutions

$$v^2 = (1-2\zeta^2) \pm 2\zeta\sqrt{1+\zeta^2} \quad (9.41)$$

If system is lightly damped, ζ^2 can be neglected and Equation 9.41 can be approximated as

$$v^2 = 1 \pm 2\zeta \quad (9.42)$$

or

$$\left(\frac{\omega_1}{\omega_n}\right)^2 = 1 - 2\zeta \quad (9.43)$$

$$\left(\frac{\omega_2}{\omega_n}\right)^2 = 1 + 2\zeta \quad (9.44)$$

Subtracting Equation 9.43 from 9.44 gives

$$\frac{\omega_2^2 - \omega_1^2}{\omega_n^2} = 4\zeta \quad (9.45)$$

In general, the natural frequency ω_n is between the half-power points and $\omega_n \approx \frac{1}{2}(\omega_1 + \omega_2)$ for light damping. Thus, we have

$$\frac{\omega_2 - \omega_1}{\omega_n} \approx 2\zeta \quad (9.46)$$

or

$$\zeta \approx \frac{\Delta\omega}{2\omega_n} \quad (9.47)$$

where $\Delta\omega = \omega_2 - \omega_1$ is referred to as the bandwidth of the system.

Example 9.3: Half-Power Bandwidth

A viscously damped system consisting of a mass of 5 kg and a spring of stiffness 12,500 N/m is subjected to a harmonic force excitation. The frequency ratios ω_1/ω_n and ω_2/ω_n at half-power points are observed to be 0.9093 and 1.0713, respectively. Assume that the system is lightly damped.

- Determine the bandwidth of the system.
- Determine the damping ratio ζ .

Solution

- Note that the undamped natural frequency is $\omega_n = \sqrt{k/m} = \sqrt{12,500/5} = 50$ rad/s. The bandwidth of the system is

$$\Delta\omega = \omega_2 - \omega_1 = \omega_n \left(\frac{\omega_2}{\omega_n} - \frac{\omega_1}{\omega_n} \right) = 50(1.0713 - 0.9093) = 8.1 \text{ rad/s}$$

- For small damping, the damping ratio is

$$\zeta \approx \frac{\Delta\omega}{2\omega_n} = \frac{8.1}{2(50)} = 0.081$$

9.2.2 Rotating Unbalance

Unbalance in rotating machines is a common source of harmonic excitation. The unbalance is caused by the rotating part, for which the mass center does not coincide with the center of rotation. [Figure 9.8](#) shows a system with an unbalanced mass m rotating at a constant angular velocity of ω . The distance between the unbalanced mass and the center of rotation is e , which represents the *eccentricity*. The mass of the entire system is M and its motion is assumed to be constrained along the vertical direction only. The entire system can therefore be considered as a single-degree-of-freedom system.

Choose the static equilibrium position of the entire system as the coordinate origin. Denote the mass of the nonrotating part as $M - m$ and the corresponding vertical displacement as the generalized coordinate x . Thus, the vertical displacement of the rotating unbalanced mass is $x + e \sin(\omega t)$. The free-body diagrams of $M - m$ and m are shown in [Figure 9.9](#).

Applying Newton's second law to $M - m$ and m along the vertical direction yields

$$-F_V - kx - b\dot{x} = (M - m)\ddot{x} \quad (9.48)$$

$$F_V = m \frac{d^2}{dt^2} [x + e \sin(\omega t)] \quad (9.49)$$

where F_V is the vertical component of the internal force between the rotating and the nonrotating parts. Combining Equations 9.48 and 9.49, and eliminating F_V , gives the differential equation of the system with rotating unbalance, as

$$(M - m)\ddot{x} + m \frac{d^2}{dt^2} [x + e \sin(\omega t)] + b\dot{x} + kx = 0 \quad (9.50)$$

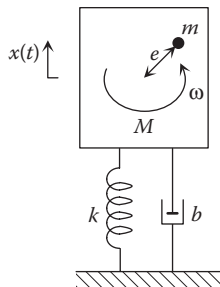


FIGURE 9.8

A system with rotating unbalance.

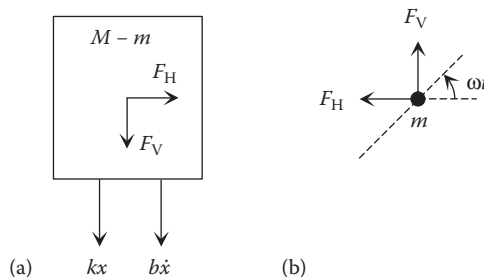


FIGURE 9.9

Free-body diagram of the system shown in [Figure 9.8](#): (a) nonrotating part $M - m$ and (b) rotating unbalance m .

Differentiating $x + e\sin(\omega t)$ with respect to time twice yields

$$M\ddot{x} + b\dot{x} + kx = me\omega^2 \sin(\omega t) \quad (9.51)$$

which implies that the effect of a rotating unbalance mass is to exert a harmonic force $me\omega^2 \sin(\omega t)$ on the system.

Note that Equation 9.51 is similar to Equation 9.29, except that the magnitude of the harmonic force is $me\omega^2$ instead of F_0 . Thus, with this modification, we can obtain the steady-state solution of the system by using Equations 9.31 through 9.33. The dynamic amplitude is given by

$$X = \frac{me\omega^2/k}{\sqrt{\left[1 - (\omega/\omega_n)^2\right]^2 + \left[2\zeta(\omega/\omega_n)\right]^2}} \quad (9.52)$$

where $\omega_n = \sqrt{k/M}$. Replacing k in Equation 9.52 with $M\omega_n^2$ yields

$$X = \frac{me(\omega/\omega_n)^2/M}{\sqrt{\left[1 - (\omega/\omega_n)^2\right]^2 + \left[2\zeta(\omega/\omega_n)\right]^2}} \quad (9.53)$$

For a system with rotating unbalance, the dimensionless ratio

$$\frac{MX}{me} = \frac{(\omega/\omega_n)^2}{\sqrt{\left[1 - (\omega/\omega_n)^2\right]^2 + \left[2\zeta(\omega/\omega_n)\right]^2}} \quad (9.54)$$

is usually plotted versus the dimensionless ratio ω/ω_n , known as normalized frequency or frequency ratio. [Figure 9.10](#) shows the magnitude of the frequency response for a system with rotating unbalance, where the vertical axis is $MX/(me)$. For low-speed rotations, that is, $\omega/\omega_n \ll 1$, the magnitude of the response $MX/(me)$ is very small, and thus the vibration amplitude X is close to zero. Resonance peaks occur in the neighborhood of $\omega/\omega_n = 1$. For high-speed rotations, that is, $\omega/\omega_n \gg 1$, the magnitude of the response $MX/(me)$ approaches 1. This implies that the dynamic amplitude X is me/M , which is constant regardless of the driving frequency and the amount of damping.

Example 9.4: Rotating Unbalance

An electric motor of mass $M = 450$ kg is mounted on a simply supported beam with negligible mass. Assume that the supporting beam is equivalent to a spring of stiffness $k = 5000$ kN/m and the damping ratio of the system is 0.1. The unbalance in the rotor of the motor is $me = 0.5$ kg·m. Determine the dynamic amplitude X of the motor when it runs at a speed of 1000 rpm.

Solution

The natural frequency of the system is

$$\omega_n = \sqrt{\frac{k}{M}} = \sqrt{\frac{5000 \times 10^3}{450}} = 105.41 \text{ rad/s}$$

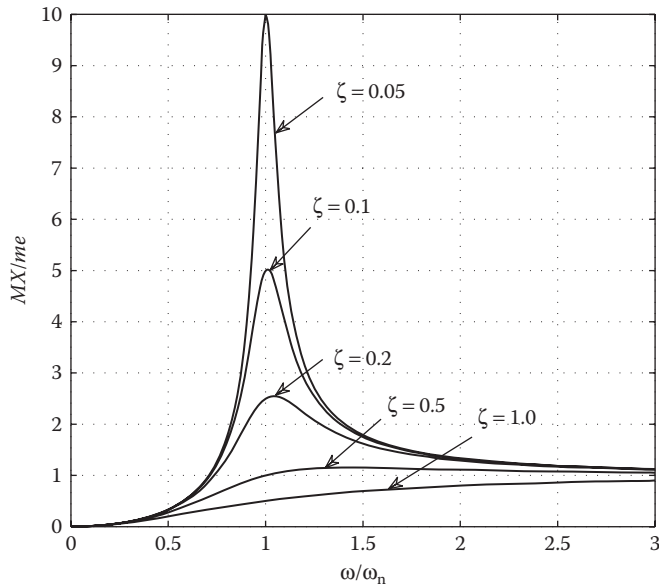


FIGURE 9.10
Magnitude of the frequency response for a system with rotating unbalance.

When the motor runs at a speed of 1000 rpm, the frequency ratio is

$$\frac{\omega}{\omega_n} = \frac{1000(2\pi)}{60(105.41)} = 0.99$$

Thus,

$$\frac{MX}{me} = \frac{(\omega/\omega_n)^2}{\sqrt{[1 - (\omega/\omega_n)^2]^2 + [2\zeta(\omega/\omega_n)]^2}} = \frac{0.99^2}{\sqrt{(1 - 0.99^2)^2 + [2(0.1)(0.99)]^2}} = 4.93$$

which gives the dynamic amplitude as

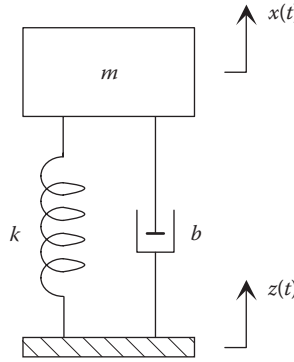
$$X = \frac{me}{M} \left(\frac{MX}{me} \right) = \frac{0.5(4.93)}{450} = 5.50 \times 10^{-3} \text{ m}$$

9.2.3 Harmonic Base Excitation

Many applications in vibration involve systems with displacement as the input. Examples include a machine placed on a foundation undergoing vibration and a vehicle traveling on a wavy road. Assume that each system mentioned can be modeled as a single-degree-of-freedom system, as shown in [Figure 9.11](#), where x and z are the displacements of the mass and the base, respectively.

Applying Newton's second law gives the differential equation of motion as

$$m\ddot{x}(t) + b\dot{x}(t) + kx(t) = b\dot{z}(t) + kz(t) \quad (9.55)$$

**FIGURE 9.11**

A single-degree-of-freedom system undergoing base excitation.

or

$$\ddot{x}(t) + 2\zeta\omega_n\dot{x}(t) + \omega_n^2x(t) = 2\zeta\omega_n\dot{z}(t) + \omega_n^2z(t) \quad (9.56)$$

As discussed in [Section 8.4](#), the system's frequency response function is

$$G(j\omega) = \frac{X(j\omega)}{Z(j\omega)} = \frac{j2\zeta\omega_n\omega + \omega_n^2}{-\omega^2 + j2\zeta\omega_n\omega + \omega_n^2} = \frac{1 + j2\zeta\omega/\omega_n}{1 - (\omega/\omega_n)^2 + j2\zeta\omega/\omega_n} \quad (9.57)$$

whose magnitude and phase are given by

$$|G(j\omega)| = \frac{\sqrt{1 + (2\zeta\omega/\omega_n)^2}}{\sqrt{[1 - (\omega/\omega_n)^2]^2 + (2\zeta\omega/\omega_n)^2}} \quad (9.58)$$

and

$$\phi = \angle G(j\omega) = -\tan^{-1} \frac{2\zeta(\omega/\omega_n)^3}{1 - (\omega/\omega_n)^2 + (2\zeta\omega/\omega_n)^2} \quad (9.59)$$

Assume that the motion of the base is harmonic; for example, $z(t) = Z_0\sin(\omega t)$. Then, the steady-state response of the system is also harmonic, $x(t) = X\sin(\omega t + \phi)$, where the dynamic amplitude X is given by

$$X = Z_0|G(j\omega)| = Z_0 \frac{\sqrt{1 + (2\zeta\omega/\omega_n)^2}}{\sqrt{[1 - (\omega/\omega_n)^2]^2 + (2\zeta\omega/\omega_n)^2}} \quad (9.60)$$

Thus, the dimensionless ratio of the dynamic amplitude X and the amplitude of the base displacement is

$$\frac{X}{Z_0} = \frac{\sqrt{1 + (2\zeta\omega/\omega_n)^2}}{\sqrt{[1 - (\omega/\omega_n)^2]^2 + (2\zeta\omega/\omega_n)^2}} \quad (9.61)$$

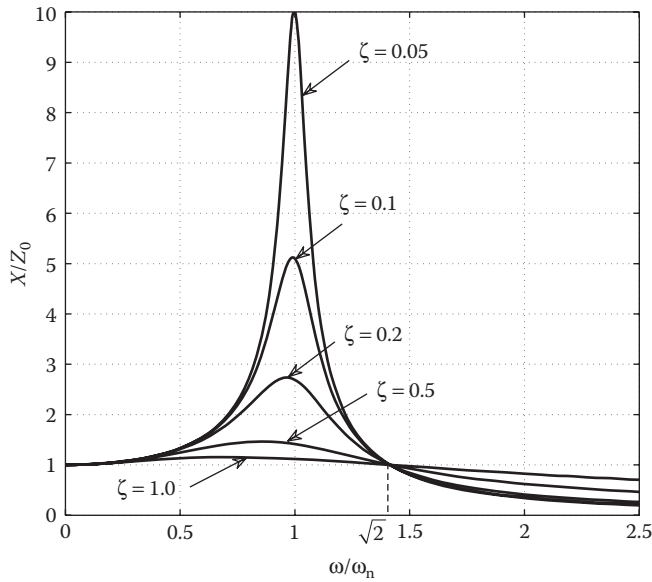


FIGURE 9.12
Magnitude of the frequency response for a system with harmonic base excitation.

which is known as the *transmissibility*. Figure 9.12 shows the magnitude of the frequency response for a system with harmonic excitation, where the vertical axis is X/Z_0 . Note that $X/Z_0 = 1$ for all curves when the frequency ratio is $\omega/\omega_n = \sqrt{2}$. This implies that the response has the same magnitude as the excitation. The response is amplified when $\omega/\omega_n < \sqrt{2}$ and is reduced when $\omega/\omega_n > \sqrt{2}$.

Example 9.5: Harmonic Base Excitation

The mass–spring–damper system in Example 5.21 represents a vehicle traveling on a rough road. Assume that the surface of the road can be approximated as a sine wave $z = Z_0 \sin(\omega t)$, where $Z_0 = 0.01$ m and $\omega = 3.5$ rad/s. The mathematical model of the system is given by an ordinary differential equation $m\ddot{x} + b\dot{x} + kx = b\dot{z} + kz$, where $m = 3000$ kg, $b = 2000$ N·s/m, and $k = 50$ kN/m. Determine the dynamic amplitude of the vehicle.

Solution

The natural frequency and the damping ratio of the vehicle are

$$\omega_n = \sqrt{\frac{k}{m}} = \sqrt{\frac{50 \times 10^3}{3000}} = 4.08 \text{ rad/s}$$

and

$$\zeta = \frac{b}{2\sqrt{mk}} = \frac{2000}{2\sqrt{(3000)(50 \times 10^3)}} = 0.08$$

The frequency ratio is

$$\frac{\omega}{\omega_n} = \frac{3.5}{4.08} = 0.86$$

and the transmissibility can be calculated via Equation 9.61, as

$$\frac{X}{Z_0} = \frac{X}{0.01} = \frac{\sqrt{1 + [2(0.08)(0.86)]^2}}{\sqrt{(1 - 0.86^2)^2 + [2(0.08)(0.86)]^2}} = 3.43$$

Thus, the dynamic amplitude of the system is $X = 3.43 \times 10^{-2}$ m.

PROBLEM SET 9.2

- Figure 9.13 shows the experimental data for the frequency response of a single-degree-of-freedom system. Use the half-power bandwidth method to estimate the damping ratio ζ of the system.
- A viscously damped single-degree-of-freedom system is subjected to a harmonic force excitation. It is observed that the amplitude of the frequency response X/x_{st} reaches the maximum when the driving frequency is 150 rpm. The peak value is 40. Assume the system to be lightly damped.
 - Determine the damping ratio ζ .
 - Determine the bandwidth of the system.
- An industrial machine of mass $M = 450$ kg is supported by a spring with a static deflection $x_{st} = 0.5$ cm. If the machine has a rotating unbalance $me = 0.25$ kg·m, determine the dynamic amplitude X at 500 rpm. Assume damping to be negligible.
- Tires must be balanced so that no periodic forces develop during operation. Figure 9.14 shows a tire with an eccentric mass because of uneven wear. The parameters are given as follows: the mass of the tire is $M = 11.75$ kg, the unbalanced

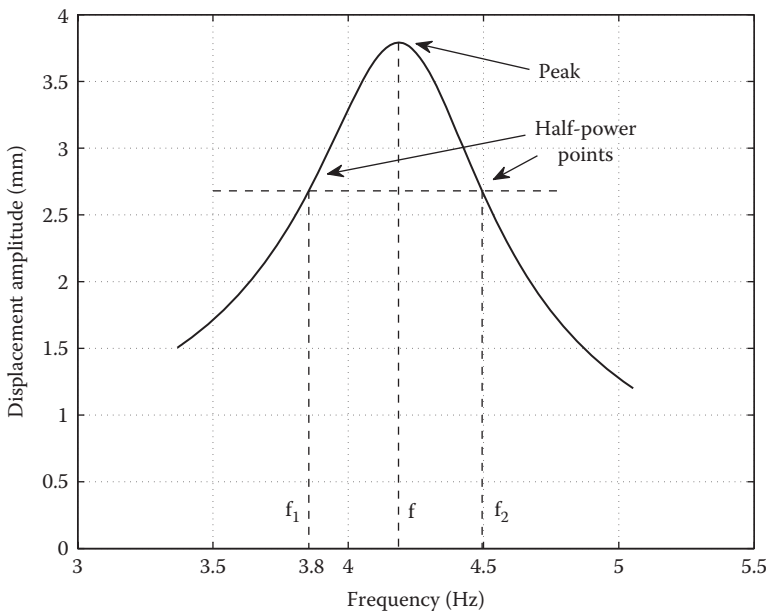
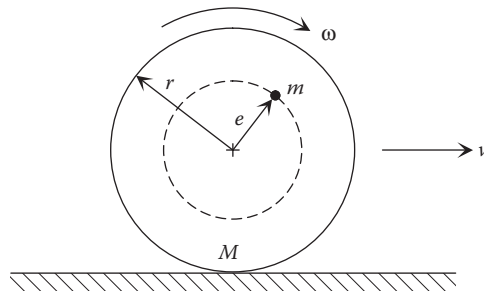


FIGURE 9.13
Problem 1.

**FIGURE 9.14**

Problem 4.

- mass is $m = 0.1$ kg, the radius of the tire is $r = 22.5$ cm, and the eccentric distance is $e = 15$ cm. Assume that the stiffness of the tire is 120 kN/m. Neglect the damping of the system. Determine the amplitude of the steady-state response of the tire caused by mass unbalance when the car moves at (a) 40 km/h and (b) 80 km/h.
5. Reconsider Example 9.5, in which the mathematical model of a vehicle is given by an ordinary differential equation $m\ddot{x} + b\dot{x} + kx = b\dot{z} + kz$ with $m = 3000$ kg, $b = 2000$ N·s/m, and $k = 50$ kN/m. It is observed that the base excitation due to the roughness of the road surface is also related to the speed of the vehicle, that is, $z = 0.01\sin(0.2\pi vt)$. Determine the transmissibility and the dynamic amplitude X of the vehicle when it moves at a speed of (a) 25 km/h and (b) 105 km/h.
 6. Precision instruments must be placed on rubber mounts, which act as springs and dampers, to reduce the effects of base vibration. Consider a precision instrument of mass 110 kg mounted on a rubber block. For the entire assembly, the spring stiffness is 250 kN/m and the damping ratio is 0.10 . Assume that the base undergoes vibration, and the displacement of the base is expressed as $y(t) = Y_0\sin(\omega t)$. Determine the dynamic amplitude of the system if the acceleration amplitude of the base excitation is 0.15 m/s² and the excitation frequency is (a) 8 Hz and (b) 24 Hz.

9.3 VibrationSuppressions

Vibrations are undesirable in most cases, particularly in cars, machining tools, precision instruments, buildings in an active seismic zone, and so on. To protect these systems and enhance their life, it is necessary to reduce vibration. This can be achieved with vibration isolators or vibration absorbers.

9.3.1 Vibration Isolators

To isolate an object from the source of vibrations, two types of vibration isolation systems are used: passive and active. Passive vibration isolation systems consist of springs and dampers. Active vibration isolation systems contain (along with the springs) piezoelectric accelerometers, electromagnetic actuators, and control circuits. The topic of active vibration isolation is beyond the scope of this text, and therefore will not be discussed.

A vibration isolation system attempts either to protect delicate equipment from vibration transmitted to it from its support system or to prevent transmission of the vibratory

force generated by a machine to its surroundings. The essence of these two objectives is the same. The concept of transmissibility introduced in [Section 9.2](#) can be used for either displacement isolation design or force isolation design.

As shown in Equation 9.61, for a system placed on a support undergoing harmonic vibration, the dynamic response depends on the natural frequency ω_n and the damping ratio ζ . This fact is also shown in [Figure 9.12](#). When the natural frequency ω_n is much less than the excitation frequency ω , more specifically, $\omega/\omega_n > \sqrt{2}$, the displacement transmissibility X/Z_0 is less than 1. This implies that the magnitude of the system response is reduced. Thus, it is desirable to design a vibration isolator such that the natural frequency of the entire assembly is within the region of $\omega/\omega_n > \sqrt{2}$. This can be achieved by placing the system on a spring-damper system. Note that the value of damping should not be too large because the displacement transmissibility increases when the damping ratio increases for $\omega/\omega_n > \sqrt{2}$.

Example 9.6: Displacement Isolation

A machine of mass 50 kg is mounted on a rubber isolator to protect it from the ground's vibration caused by the operation of other machines nearby. Assume that the ground vibrates at 10 Hz. Determine the stiffness of the rubber isolation spring if only 20% of the ground's motion is transmitted to the machine.

- Neglect damping.
- Assume that the damping coefficient of the rubber isolator is 0.1.

Solution

- If only 20% of the ground's motion is transmitted to the machine, then the transmissibility given by Equation 9.61 with damping neglected is

$$\frac{X}{Z_0} = \frac{1}{|1 - (\omega/\omega_n)^2|} = 0.20$$

from which the frequency ratio ω/ω_n is found to be $\sqrt{6}$. For the excitation frequency $\omega = 10$ Hz, the natural frequency is

$$\omega_n = \frac{2\pi(10)}{\sqrt{6}} = 25.65 \text{ rad/s}$$

Thus, the stiffness of the rubber isolation spring is

$$k = \omega_n^2 m = 25.65^2 (50) \approx 32.90 \text{ kN/m}$$

- If the damping coefficient of the rubber isolator is 0.1, then the transmissibility is

$$\frac{X}{Z_0} = \frac{\sqrt{1 + [2(0.1)\omega/\omega_n]^2}}{\sqrt{[1 - (\omega/\omega_n)^2]^2 + [2(0.1)\omega/\omega_n]^2}} = 0.20$$

which can be rewritten as

$$(\omega/\omega_n)^4 - 2.96(\omega/\omega_n)^2 - 24 = 0$$

Solving for $(\omega/\omega_n)^2$ gives

$$(\omega/\omega_n)^2 = 6.60 \quad \text{or} \quad (\omega/\omega_n)^2 = -3.64$$

The negative value is not valid. Thus, the frequency ratio ω/ω_n is $\sqrt{6.60}$. Repeating the calculation in Part (a) gives $k \approx 29.92 \text{ kN/m}$.

For a machine placed on a rigid base, the machine itself is a vibration source. The vibratory force generated by the machine will be transmitted to the base and affect the surrounding equipment. To reduce the damaging effect of the vibratory machine on its surroundings, it is necessary to isolate it from the base.

Figure 9.15a shows a machine placed on a rigid foundation through a spring and damper system. The machine is subjected to a harmonic excitation force $f(t)$. Recall that the displacement response $x(t)$ to harmonic excitation is given by Equations 9.31 through 9.33. Differentiating Equation 9.31 with respect to time yields the velocity response, as given in Equation 9.37, in which X is the amplitude of the displacement, and the amplitude of the velocity is ωX . Moreover, the velocity leads the displacement by the phase angle $\pi/2$, because $\cos(\omega t + \phi)$ in Equation 9.37 can be expressed as $\sin(\omega t + \phi + \pi/2)$.

From the free-body diagram in Figure 9.15b, it is clear that the force transmitted to the base includes two parts: the spring force kx and the damping force $b\dot{x}$. The amplitudes of the spring force and the damping force are kX and $b\omega X$, respectively. The force vectors are shown in Figure 9.15c. The angle between the two force vectors is $\pi/2$. Thus, the amplitude of the force transmitted to the base is

$$F_T = \sqrt{(kX)^2 + (b\omega X)^2} \quad (9.62)$$

Note that $b/k = (b/m)(m/k) = 2\zeta\omega_n/\omega_n^2 = 2\zeta/\omega_n$. Then, Equation 9.62 can be rewritten as

$$F_T = kX\sqrt{1 + (2\zeta\omega/\omega_n)^2} \quad (9.63)$$

Substituting Equation 9.32 into Equation 9.63 yields

$$F_T = F_0 \frac{\sqrt{1 + (2\zeta\omega/\omega_n)^2}}{\sqrt{[1 - (\omega/\omega_n)^2]^2 + [2\zeta(\omega/\omega_n)]^2}} \quad (9.64)$$

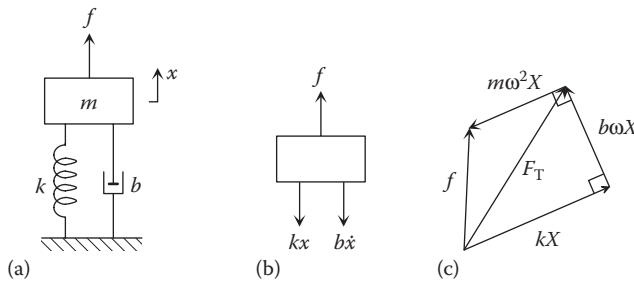


FIGURE 9.15

Force isolation: (a) physical system, (b) free-body diagram, and (c) force vector diagram.

The dimensionless ratio F_T/F_0 is the transmissibility given by Equation 9.61, and is a measure of the force transmitted to the base. Thus, the plot of F_T/F_0 versus ω/ω_n is the same as the plot of X/Z_0 versus ω/ω_n , as in Figure 9.12. When the frequency ratio is $\omega/\omega_n = \sqrt{2}$, $F_T/F_0 = 1$ and the excitation force is fully transmitted to the base. For $\omega/\omega_n > \sqrt{2}$, the force transmissibility F_T/F_0 is less than 1 and the force transmitted reduces with increasing excitation frequency ω .

Example 9.7: Force Isolation

A rotating machine of mass 2000 kg is mounted on an isolator block of stiffness 500 kN/m and damping ratio $\zeta = 0.1$. The machine is subjected to a harmonic disturbance force, for which the frequency is the same as the rotational speed of the machine. Assuming that 10% of the disturbance force is transmitted to the base, determine the rotational speed of the machine.

Solution

From Equation 9.64, we have

$$\frac{F_T}{F_0} = \frac{\sqrt{1 + [2(0.1)(\omega/\omega_n)]^2}}{\sqrt{[1 - (\omega/\omega_n)^2]^2 + [2(0.1)(\omega/\omega_n)]^2}} = 0.1$$

Solving for the frequency ratio yields $\omega/\omega_n = 3.40$. The natural frequency of the system is $\omega_n = \sqrt{500,000/2000} = 15.81$ rad/s. Thus, the excitation frequency is

$$\omega = 3.40(15.81) = 53.75 \text{ rad/s}$$

which is also the rotational speed of the machine.

9.3.2 Vibration Absorbers

As discussed in Section 8.4, for a single-degree-of-freedom system subjected to harmonic excitation, violent vibration is induced when the excitation frequency is close to the natural frequency of the system. To protect the system, we can change either the mass or the spring stiffness so that the natural frequency is not too close to the excitation frequency. However, this may not always be possible. To circumvent this issue, a vibration absorber consisting of a second mass and spring can be added to the system to protect the original single-degree-of-freedom system from harmonic excitation.

Consider the two-degree-of-freedom system shown in Figure 9.16, where m_1 and k_1 are the mass and the spring stiffness of the primary system, and m_2 and k_2 are the mass and the spring stiffness of the absorber. A harmonic force $F_1 \sin(\omega t)$ is applied to the primary system, which would undergo violent vibrations if the absorber is not installed.

For the combined system, the equations of motion in matrix form are

$$\begin{bmatrix} m_1 & 0 \\ 0 & m_2 \end{bmatrix} \begin{Bmatrix} \ddot{x}_1 \\ \ddot{x}_2 \end{Bmatrix} + \begin{bmatrix} k_1 + k_2 & -k_2 \\ -k_2 & k_2 \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix} = \begin{Bmatrix} F_1 \\ 0 \end{Bmatrix} \sin(\omega t) \quad (9.65)$$

Because the system is undamped, the steady-state response to a sinusoidal input is still sinusoidal, which has the same frequency as the excitation frequency and a zero phase angle. We can express the steady-state response as

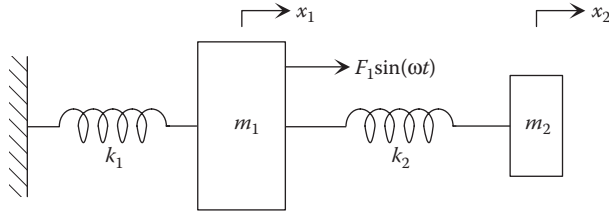


FIGURE 9.16
A vibration absorber.

$$\begin{Bmatrix} x_1(t) \\ x_2(t) \end{Bmatrix} = \begin{Bmatrix} X_1 \\ X_2 \end{Bmatrix} \sin(\omega t) \quad (9.66)$$

Substituting Equation 9.66 into Equation 9.65 yields

$$\begin{bmatrix} k_1 + k_2 - m_1\omega^2 & -k_2 \\ -k_2 & k_2 - m_2\omega^2 \end{bmatrix} \begin{Bmatrix} X_1 \\ X_2 \end{Bmatrix} = \begin{Bmatrix} F_1 \\ 0 \end{Bmatrix} \quad (9.67)$$

Solving for X_1 and X_2 gives

$$\begin{Bmatrix} X_1 \\ X_2 \end{Bmatrix} = \frac{1}{\Delta(\omega)} \begin{bmatrix} k_2 - m_2\omega^2 & k_2 \\ k_2 & k_1 + k_2 - m_1\omega^2 \end{bmatrix} \begin{Bmatrix} F_1 \\ 0 \end{Bmatrix} \quad (9.68)$$

where

$$\Delta(\omega) = (k_2 - m_2\omega^2)(k_1 + k_2 - m_1\omega^2) - k_2^2 \quad (9.69)$$

Thus, the amplitudes are

$$X_1 = \frac{(k_2 - m_2\omega^2)F_1}{(k_2 - m_2\omega^2)(k_1 + k_2 - m_1\omega^2) - k_2^2} \quad (9.70)$$

$$X_2 = \frac{k_2 F_1}{(k_2 - m_2\omega^2)(k_1 + k_2 - m_1\omega^2) - k_2^2} \quad (9.71)$$

Dividing the numerators and denominators in Equations 9.70 and 9.71 by $k_1 k_2$, we obtain

$$X_1 = \frac{[1 - (m_2/k_2)\omega^2](F_1/k_1)}{[1 - (m_2/k_2)\omega^2][1 + k_2/k_1 - (m_1/k_1)\omega^2] - (k_2/k_1)} \quad (9.72)$$

$$X_2 = \frac{F_1/k_1}{[1 - (m_2/k_2)\omega^2][1 + k_2/k_1 - (m_1/k_1)\omega^2] - (k_2/k_1)} \quad (9.73)$$

From Equation 9.72, it is clear that the amplitude X_1 of the primary system reduces to zero when the excitation frequency ω equals $\sqrt{k_2/m_2}$. Correspondingly, the amplitude X_2 of the absorber becomes

$$X_2 = -\frac{F_1}{k_2} \quad (9.74)$$

where the negative sign indicates that the absorber moves in the opposite direction of the force F_1 . Thus, the force exerted on the primary mass m_1 by the spring of the absorber is

$$k_2 x_2 = k_2 X_2 \sin \omega t = -F_1 \sin \omega t \quad (9.75)$$

which exactly balances the externally applied force $F_1 \sin(\omega t)$.

Let us denote the natural frequency of the original single-degree-of-freedom system alone as $\omega_1 = \sqrt{k_1/m_1}$ and the natural frequency of the absorber alone as $\omega_2 = \sqrt{k_2/m_2}$. To determine the parameters of the absorber, we introduce the following notations: $x_{st} = F_1/k_1$, $\mu = m_2/m_1$, and $v = \omega_2/\omega_1$, where x_{st} is the static deflection of the original system, μ is the mass ratio, and v is the natural frequency ratio. Then, Equations 9.72 and 9.73 can be rewritten as

$$X_1 = \frac{[1 - (\omega/\omega_2)^2]x_{st}}{[1 - (\omega/\omega_2)^2][1 + k_2/k_1 - (\omega/\omega_1)^2] - (k_2/k_1)} \quad (9.76)$$

$$X_2 = \frac{x_{st}}{[1 - (\omega/\omega_2)^2][1 + k_2/k_1 - (\omega/\omega_1)^2] - (k_2/k_1)} \quad (9.77)$$

Note that $\omega/\omega_2 = (\omega/\omega_1)(1/v)$ and $k_2/k_1 = (m_2/m_1)(k_2/m_2)(m_1/k_1) = \mu v^2$. Thus,

$$X_1 = \frac{[v^2 - (\omega/\omega_1)^2]x_{st}}{[v^2 - (\omega/\omega_1)^2][1 + \mu v^2 - (\omega/\omega_1)^2] - \mu v^4} \quad (9.78)$$

$$X_2 = \frac{v^2 x_{st}}{[v^2 - (\omega/\omega_1)^2][1 + \mu v^2 - (\omega/\omega_1)^2] - \mu v^4} \quad (9.79)$$

From Equations 9.78 and 9.79, it is clear that the values of X_1 and X_2 depend on the mass ratio μ and the natural frequency ratio v . Note that the natural frequency ratio v must be very close to 1. Adding the absorber aims to alleviate the vibration of the original system at resonance, that is, $\omega = \omega_1$, and the absorber can only be effective when its own natural frequency is the same as the excitation frequency, that is, $\omega_2 = \omega$. Thus, we have $\omega_2 = \omega_1$ or $v = 1$. If v is very close to 1, the motion of m_1 is not zero, but its amplitude is still very small.

Figures 9.17 shows two curves, X_1/x_{st} versus ω/ω_1 (Figure 9.17a) and X_2/x_{st} versus ω/ω_1 (Figure 9.17b) for $\mu = 0.2$ and $v = 1$. Note that the horizontal axis can be replaced by ω/ω_2 because $v = 1$, that is, $\omega_2 = \omega_1$. As observed from Figure 9.17a, when the excitation frequency

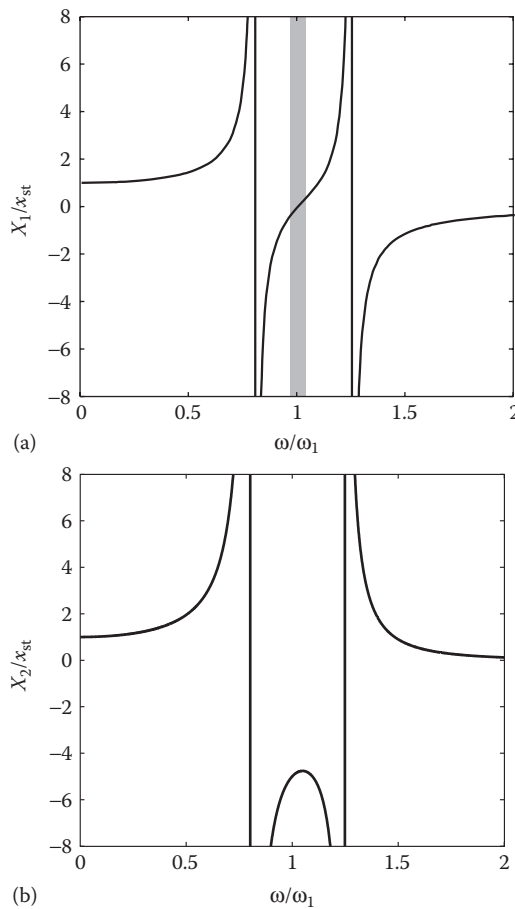
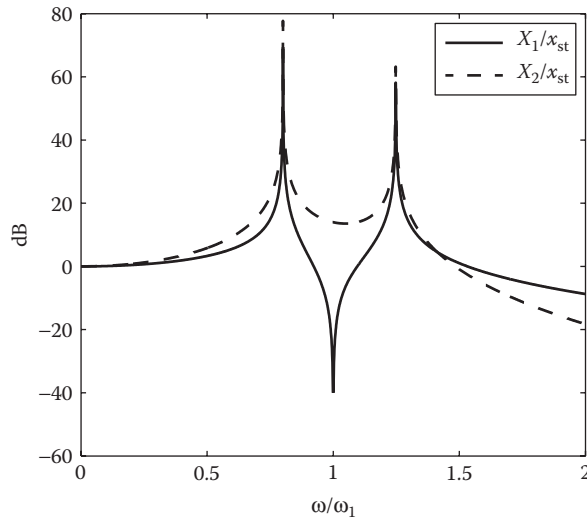


FIGURE 9.17
Frequency response curves: (a) main mass and (b) absorber mass.

ω is in the immediate neighborhood of ω_2 , the amplitude X_1 is very small. However, when the excitation frequency ω shifts slightly away from ω_2 , the amplitude X_1 increases significantly. This shows that the absorber is only effective when the excitation frequency ω is close to ω_2 .

To design a vibration absorber, we first need to select the operation frequency ω , at which the displacement amplitude of the primary mass will be tuned to zero. Then the relation between the mass and the spring stiffness of the absorber is obtained by $\omega^2 = \omega_2^2 = k_2/m_2$. Select appropriate values for m_2 and k_2 by considering restrictions on the motion of the absorber mass. Once the absorber is designed, the mass ratio μ is checked, for which the recommended value is $\mu < 0.25$.

It should be pointed out that one disadvantage of the vibration absorber is that two new resonant frequencies in the neighborhood of the excitation frequency are created as seen from Figure 9.18. Because the values of X_1/x_{st} and X_2/x_{st} change dramatically, the vertical

**FIGURE 9.18**

Frequency response curves (in dB) for the main mass and the absorber mass.

axis in Figure 9.18 is given in the decibel units, which can be used to conveniently represent very large or small numbers. The details on how to determine natural frequencies for multi-degree-of-freedom systems will be discussed in the next section.

Example 9.8: Vibration Absorber Design

When an engine operates at a speed of 4500 rpm, vibration is induced through its pedestal mount. The amplitude of the excitation force is 300 N. Design a vibration absorber that will reduce the vibration when mounted on the pedestal. Assume that the maximum allowable displacement of the absorber is 2 mm.

Solution

The natural frequency of the absorber is required to be the same as the excitation frequency,

$$\omega_2 = \omega = \frac{4500(2\pi)}{60} = 150\pi \text{ rad/s}$$

The maximum allowable displacement amplitude of the absorber is 2 mm. From Equation 9.74, considering the magnitude only, the spring stiffness of the absorber can be obtained as

$$k_2 = \frac{F_1}{X_2} = \frac{300}{2 \times 10^{-3}} = 150000 \text{ N/m}$$

Thus, the mass of the absorber is

$$m_2 = \frac{k_2}{\omega_2^2} = \frac{150000}{(150\pi)^2} = 0.68 \text{ kg}$$

PROBLEM SET 9.3

1. A 25-kg instrument is suspended from a ceiling by four springs, each of which has a stiffness of 800 N/m. The ceiling vibrates with a frequency of 2 Hz and amplitude of 0.05 mm due to the air conditioning compressor and chiller mounted on the roof. Neglecting the damping, determine the maximum displacement amplitude of the instrument.
2. Consider a 10-kg instrument placed on a floor that vibrates with a frequency of 3000 rpm and amplitude of 2 mm due to nearby machinery. A vibration isolator is designed to protect the instrument from the vibration of the floor. Assume that the damping ratio of the isolator is 0.05 and the maximum allowable acceleration amplitude of the instrument is $2g$, where g is the gravitational acceleration,
 - a. Determine the stiffness of the isolator.
 - b. Determine the maximum displacement amplitude of the instrument.
3. A 9000-kg air conditioning compressor mounted on a roof is supported by four springs. The static deformation of each spring is 4 cm. Assuming damping is negligible, find the force transmissibility when the compressor works at 60 Hz.
4. Consider the single-degree-of-freedom system shown in [Figure 9.8](#). The excitation force due to the rotating unbalance is $me\omega^2\sin(\omega t)$. Assume that the system has the following parameters: $m = 5$ kg, $M = 100$ kg, $e = 0.1$ m, $k = 5,000$ N/m, and $b = 200$ N·s/m.
 - a. Determine the force transmitted to the support when the system runs at the rotating speed $\omega = 1.5\omega_n$.
 - b. Determine the force transmissibility.
 - c. The spring is replaced to decrease the force transmissibility at the same excitation frequency to 20%. Determine the stiffness of the new spring.
5. A rotating machine has a mass of 6 kg and a natural frequency of 5 Hz. Due to a rotating unbalanced mass, the machine is subjected to a harmonic disturbance force $me\omega^2\sin(\omega t)$. When the machine operates at a frequency of 3.5 Hz, the amplitude of the disturbance force is 40 N.
 - a. Design a vibration absorber assuming that the maximum allowable displacement of the absorber is 5 cm.
 - b.  Using MATLAB®, write an m-file to plot X_1/me versus ω . Use the decibel units to show the vertical axis.
6. The pendulum in [Figure 9.19](#) is known as a tuned mass damper. It is mounted in a building, which is simplified as a block of mass m_1 supported by a spring of stiffness k . The mass of the pendulum is m_2 and the length l is tunable. The tunable pendulum is used to control the vibration of the building under extreme wind loads. Assume that the force due to gusty winds can be modeled as a harmonic force $F_1\sin(\omega t)$, and the excitation frequency ω is very close to the natural frequency of the building. Prove that the forced vibration of the building can be eliminated when the length of the pendulum is tuned such that $\omega = \sqrt{g/l}$, where g is the gravitational acceleration.

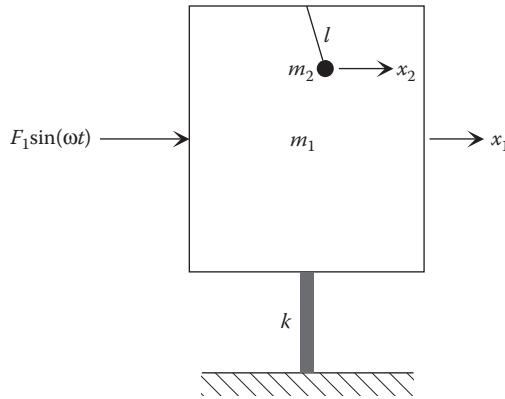


FIGURE 9.19
Problem 6.

9.4 Modal Analysis

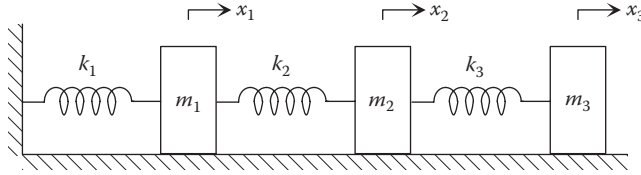
The systems discussed in the previous sections are mainly single-degree-of-freedom systems, for which the vibration can be studied using elementary methods presented in [Chapter 8](#). However, for a multi-degree-of-freedom system, more advanced mathematical tools are required to solve the equations of motion due to coordinate coupling. In this section, we first introduce key concepts, such as the eigenvalue problem, natural modes, and orthogonality of modes. Then, we develop modal analysis in a rigorous manner to decouple coordinates and use it to obtain the response to initial excitations or external forces. All derivations in this section are presented in matrix form.

9.4.1 Eigenvalue Problem

Consider a three-degree-of-freedom mass–spring system, as shown in [Figure 9.20](#), where the motion of the system is described by the displacement coordinates x_1 , x_2 , and x_3 . In the absence of damping and external forces, the system undergoes undamped free vibration, and the differential equations of motion in matrix form are

$$\begin{bmatrix} m_1 & 0 & 0 \\ 0 & m_2 & 0 \\ 0 & 0 & m_3 \end{bmatrix} \begin{Bmatrix} \ddot{x}_1 \\ \ddot{x}_2 \\ \ddot{x}_3 \end{Bmatrix} + \begin{bmatrix} k_1 + k_2 & -k_2 & 0 \\ -k_2 & k_2 + k_3 & -k_3 \\ 0 & -k_3 & k_3 \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \\ x_3 \end{Bmatrix} = \begin{Bmatrix} 0 \\ 0 \\ 0 \end{Bmatrix} \quad (9.80)$$

Note that the mass matrix is diagonal but the stiffness matrix is not. Thus, Equation 9.80 represents a set of three simultaneous (or coupled) second-order differential equations. It is generally difficult to obtain the analytical solution of Equation 9.80 because of coordinate coupling. If we can find a coordinate transformation that simultaneously diagonalizes the mass and stiffness matrices, then the system dynamics can be decoupled into a set of independent second-order differential equations, each in the form of $m\ddot{x} + kx = 0$, which can be easily solved as a single-degree-of-freedom system. Such coordinate transformation, which is not unique, indeed exists and may be found by solving the eigenvalue problem.

**FIGURE 9.20**

A three-degree-of-freedom mass–spring system.

The eigenvalue problem is a problem associated with free, undamped vibration, for which the differential equation in the general matrix form is given by

$$\mathbf{M}\ddot{\mathbf{x}} + \mathbf{K}\mathbf{x} = \mathbf{0} \quad (9.81)$$

The solution of Equation 9.81 is harmonic. Assume that the solution is in the form

$$\mathbf{x} = \mathbf{X}e^{j\omega t} \quad (9.82)$$

where ω is the frequency of the harmonic motion. Note that $\ddot{\mathbf{x}} = -\omega^2 \mathbf{X}e^{j\omega t}$. Thus, Equation 9.81 becomes

$$(\mathbf{K} - \omega^2 \mathbf{M})\mathbf{X}e^{j\omega t} = \mathbf{0} \quad (9.83)$$

Canceling the non-zero term $e^{j\omega t}$ leads to

$$(\mathbf{K} - \omega^2 \mathbf{M})\mathbf{X} = \mathbf{0} \quad (9.84)$$

or

$$\mathbf{K}\mathbf{X} = \omega^2 \mathbf{M}\mathbf{X} \quad (9.85)$$

Either one of Equations 9.84 or 9.85 represents the eigenvalue problem associated with matrices $\mathbf{M}$ and $\mathbf{K}$. In particular, it is known as the algebraic eigenvalue problem, whose solution process is similar to that of the eigenvalue problem associated with a matrix $\mathbf{A}$, as discussed in [Section 3.3](#). Equation 9.84 has a nontrivial solution $\mathbf{X} \neq \mathbf{0}$ if and only if the coefficient matrix is singular,

$$|\mathbf{K} - \omega^2 \mathbf{M}| = 0 \quad (9.86)$$

Equation 9.86 is known as the characteristic equation or *frequency equation*. For an n -degree-of-freedom system, the determinant $|\mathbf{K} - \omega^2 \mathbf{M}|$ is a polynomial of degree n in ω^2 . The n roots of Equation 9.86 are referred to as eigenvalues and denoted by $\omega_1^2, \omega_2^2, \dots$, and ω_n^2 . Once the eigenvalues are identified, each eigenvector corresponding to each of the eigenvalues can be obtained by solving Equation 9.84 or 9.85. The n eigenvectors are denoted by $\mathbf{X}_1, \mathbf{X}_2, \dots$, and $\mathbf{X}_n$.

For a vibration system, the eigenvalues and eigenvectors associated with the eigenvalue problem defined in Equation 9.84 or 9.85 have significant physical meanings. The square roots of the eigenvalues are the system's natural frequencies ω_r , where $r = 1, 2, \dots, n$. The natural frequencies are usually arranged in increasing order of magnitude, that is, $\omega_1 \leq \omega_2 \leq \dots \leq \omega_n$. The lowest frequency ω_1 is referred to as the fundamental frequency, which is extremely important for many practical problems. The eigenvectors are referred

to as modal vectors. Each modal vector represents physically the shape of a normal mode, a certain pattern of motion in which all masses move harmonically with the same natural frequency associated with this modal vector. The following example shows how to solve the eigenvalue problem and describes the physical significance of the eigenvalues and eigenvectors.

Example 9.9: Natural Frequencies and Modal Vectors

Consider the three-degree-of-freedom system shown in Figure 9.20. Assume that $m_1 = m_2 = m_3 = m$, $k_1 = 3k$, $k_2 = 2k$, and $k_3 = k$.

- Solve the associated eigenvalue problem by hand.
- Assuming $m = 1$ kg and $k = 1$ N/m, solve the associated eigenvalue problem in MATLAB.

Solution

- The differential equations for the system in Figure 9.20 are given by Equation 9.80. Thus, the mass and stiffness matrices are

$$\mathbf{M} = \begin{bmatrix} m & 0 & 0 \\ 0 & m & 0 \\ 0 & 0 & m \end{bmatrix}, \quad \mathbf{K} = \begin{bmatrix} 5k & -2k & 0 \\ -2k & 3k & -k \\ 0 & -k & k \end{bmatrix}$$

Substituting $\mathbf{M}$ and $\mathbf{K}$ into the frequency equation defined by Equation 9.86, we have

$$\begin{vmatrix} 5k - \omega^2 m & -2k & 0 \\ -2k & 3k - \omega^2 m & -k \\ 0 & -k & k - \omega^2 m \end{vmatrix} = 0$$

To simplify the calculation, divide the above-mentioned equation by k and let $\beta = \omega^2 m/k$. Thus, the determinant is

$$\begin{vmatrix} 5 - \beta & -2 & 0 \\ -2 & 3 - \beta & -1 \\ 0 & -1 & 1 - \beta \end{vmatrix} = (5 - \beta)[(3 - \beta)(1 - \beta) - 1] - (-2)[(-2)(1 - \beta)] = 0$$

which reduces to

$$\beta^3 - 9\beta^2 + 18\beta - 6 = 0$$

The three roots are

$$\beta_1 = 0.4158, \quad \beta_2 = 2.2943, \quad \beta_3 = 6.2899$$

which give the three eigenvalues

$$\omega_1^2 = 0.4158 \frac{k}{m}, \quad \omega_2^2 = 2.2943 \frac{k}{m}, \quad \omega_3^2 = 6.2899 \frac{k}{m}$$

Thus, the three natural frequencies are

$$\omega_1 = 0.6448\sqrt{\frac{k}{m}}, \quad \omega_2 = 1.5147\sqrt{\frac{k}{m}}, \quad \omega_3 = 2.5080\sqrt{\frac{k}{m}}$$

To determine the eigenvectors or modal vectors, we insert $\omega_r^2 (r = 1, 2, 3)$ into Equation 9.84. For $r = 1$, we have

$$(\mathbf{K} - \omega_1^2 \mathbf{M})\mathbf{X}_1 = \mathbf{0}$$

Note that the modal vectors are 3×1 column vectors for a three-degree-of-freedom system. The above-mentioned equation can be written as

$$\begin{bmatrix} 5k - \omega_1^2 m & -2k & 0 \\ -2k & 3k - \omega_1^2 m & -k \\ 0 & -k & k - \omega_1^2 m \end{bmatrix} \begin{bmatrix} X_{11} \\ X_{12} \\ X_{13} \end{bmatrix} = \begin{bmatrix} 0 \\ 0 \\ 0 \end{bmatrix}$$

Again, to simplify the calculation, we let β_1 replace $\omega_1^2 m/k$ and write the equations in scalar form

$$\begin{aligned} (5 - \beta_1)X_{11} - 2X_{12} &= 0 \\ -2X_{11} + (3 - \beta_1)X_{12} - X_{13} &= 0 \\ -X_{12} + (1 - \beta_1)X_{13} &= 0 \end{aligned}$$

Note that combining the first and the third equations by canceling X_{12} gives

$$\frac{X_{13}}{X_{11}} = \frac{5 - \beta_1}{2 - 2\beta_1}$$

Then, expressing X_{13} in the second equation in terms of X_{11} yields

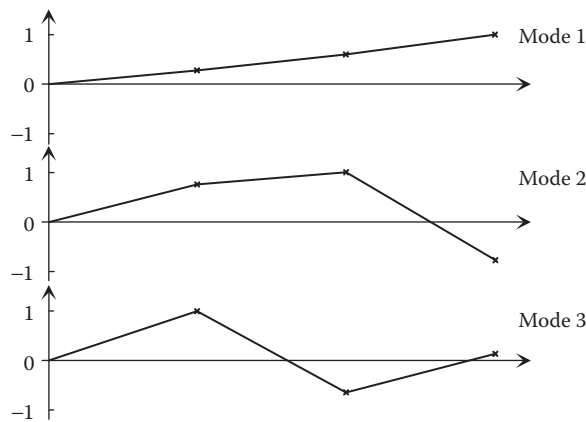
$$\frac{X_{12}}{X_{11}} = \frac{9 - 5\beta_1}{(2 - 2\beta_1)(3 - \beta_1)}$$

If we assign an arbitrary value to one element in $\mathbf{X}_1$, the other two can be determined uniquely. This implies that we have three equations and two unknowns. Assuming $X_{11} = 1$, we find $X_{13} = 3.9235$ and $X_{12} = 2.2922$. Thus, the first modal vector is

$$\mathbf{X}_1 = \begin{bmatrix} 1 \\ 2.2922 \\ 3.9235 \end{bmatrix}$$

Similarly, we can find the other two modal vectors as

$$\mathbf{X}_2 = \begin{bmatrix} 1 \\ 1.3529 \\ -1.0452 \end{bmatrix}, \quad \mathbf{X}_3 = \begin{bmatrix} 1 \\ -0.6450 \\ 0.1219 \end{bmatrix}$$

**FIGURE 9.21**

Natural modes of the three-degree-of-freedom system shown in Figure 9.20.

Here, we set the magnitude of the first component in all modal vectors as 1. The modal vectors can also be normalized so that the largest component in magnitude is equal to 1:

$$\mathbf{X}_1 = \begin{bmatrix} 0.2549 \\ 0.5842 \\ 1 \end{bmatrix}, \quad \mathbf{X}_2 = \begin{bmatrix} 0.7392 \\ 1 \\ -0.7726 \end{bmatrix}, \quad \mathbf{X}_3 = \begin{bmatrix} 1 \\ -0.6450 \\ 0.1219 \end{bmatrix}$$

The modal vectors represent physically the shape of the natural modes, as shown in Figure 9.21. In mode 1, all masses move in the same direction and oscillate with a frequency of $\omega_1 = 0.6448\sqrt{k/m}$. In mode 2, all masses oscillate with a frequency of $\omega_2 = 1.5147\sqrt{k/m}$. Masses 1 and 2 move in the same direction, while mass 3 moves in the opposite direction. The direction change is also implied by the one sign change in the second mode $\mathbf{X}_2$. In mode 3, there are two sign changes in $\mathbf{X}_3$. Masses 1 and 3 move in the same direction, while mass 2 moves in the opposite direction. All masses oscillate with a frequency of $\omega_3 = 2.5080\sqrt{k/m}$.

- b.  Given the mass and stiffness matrices, it is easy to use MATLAB to solve the associated algebraic eigenvalue problem. The following is the MATLAB session.

```
>> M = eye(3);
>> K = [5 -2 0; -2 3 -1; 0 -1 1];
>> [V,D] = eig(K,M);
```

The `eig` command returns two matrices

$$\mathbf{V} = \begin{bmatrix} 0.2149 & -0.5049 & -0.8360 \\ 0.4927 & -0.6831 & 0.5392 \\ 0.8433 & 0.5277 & -0.1019 \end{bmatrix}, \quad \mathbf{D} = \begin{bmatrix} 0.4158 & 0 & 0 \\ 0 & 2.2943 & 0 \\ 0 & 0 & 6.2899 \end{bmatrix}$$

where:

$\mathbf{D}$ is a diagonal matrix containing the eigenvalues
columns in matrix $\mathbf{V}$ are the corresponding eigenvectors

Note that the modal vectors returned by the MATLAB command `eig` are normalized so that the magnitude of each vector is equal to 1. For example, the first column in matrix $\mathbf{V}$ is the first modal vector, and we have

$$\sqrt{0.2149^2 + 0.4927^2 + 0.8433^2} = 1$$

Comparing the eigenvectors obtained by hand and by MATLAB, we observe that the ratios between the elements are the same. For example, for mode 1, $X_{11}/X_{13} = 0.2149/0.8433 = 0.2548$ and $X_{12}/X_{13} = 0.4927/0.8433 = 0.5843$. This implies that the shape of modes is unique but the amplitude is not. Due to round-off error, the ratios are slightly different from that obtained in Part (a).

9.4.2 Orthogonality of Modes

The natural modes possess an important property known as orthogonality. To show this concept, let us consider an n -degree-of-freedom system. Following the same procedure as in Example 9.9, we can obtain n natural frequencies and n modal vectors by solving the eigenvalue problem. Denote any two pairs of solutions of the eigenvalue problem as ω_r^2 , $\mathbf{X}_r$ and ω_s^2 , $\mathbf{X}_s$. Both pairs should satisfy Equation 9.85, that is,

$$\mathbf{K}\mathbf{X}_r = \omega_r^2 \mathbf{M}\mathbf{X}_r \quad (9.87)$$

$$\mathbf{K}\mathbf{X}_s = \omega_s^2 \mathbf{M}\mathbf{X}_s \quad (9.88)$$

Pre-multiplying both sides of Equation 9.87 by $\mathbf{X}_s^T$ and Equation 9.88 by $\mathbf{X}_r^T$ gives

$$\mathbf{X}_s^T \mathbf{K}\mathbf{X}_r = \omega_r^2 \mathbf{X}_s^T \mathbf{M}\mathbf{X}_r \quad (9.89)$$

$$\mathbf{X}_r^T \mathbf{K}\mathbf{X}_s = \omega_s^2 \mathbf{X}_r^T \mathbf{M}\mathbf{X}_s \quad (9.90)$$

Taking the transpose of Equation 9.90 on both sides yields

$$\mathbf{X}_s^T \mathbf{K}^T \mathbf{X}_r = \omega_s^2 \mathbf{X}_s^T \mathbf{M}^T \mathbf{X}_r \quad (9.91)$$

Recall that the mass and stiffness matrices are symmetric, $\mathbf{M} = \mathbf{M}^T$ and $\mathbf{K} = \mathbf{K}^T$. Then, Equation 9.91 reduces to

$$\mathbf{X}_s^T \mathbf{K}\mathbf{X}_r = \omega_s^2 \mathbf{X}_s^T \mathbf{M}\mathbf{X}_r \quad (9.92)$$

Subtracting Equation 9.92 from Equation 9.89, we have

$$(\omega_r^2 - \omega_s^2) \mathbf{X}_s^T \mathbf{M}\mathbf{X}_r = 0 \quad (9.93)$$

For two distinct modes, the frequencies are different, $\omega_r^2 \neq \omega_s^2$. Thus, Equation 9.93 is satisfied if and only if

$$\mathbf{X}_s^T \mathbf{M}\mathbf{X}_r = 0, \quad r \neq s \quad (9.94)$$

Substituting Equation 9.94 into Equation 9.89 gives

$$\mathbf{X}_s^T \mathbf{K}\mathbf{X}_r = 0, \quad r \neq s \quad (9.95)$$

Equations 9.94 and 9.95 represent the orthogonality relation for two distinct modal vectors, $\mathbf{X}_r$ and $\mathbf{X}_s$. They are orthogonal with respect to the mass matrix $\mathbf{M}$ as well as the stiffness matrix $\mathbf{K}$. If $r = s$, then the values of $\mathbf{X}_r^T \mathbf{M} \mathbf{X}_r$ and $\mathbf{X}_r^T \mathbf{K} \mathbf{X}_r$ are non-zero, and their values depend on the method used to normalize the modal vectors. A convenient scheme is to normalize $\mathbf{X}_r$ such that

$$\mathbf{X}_r^T \mathbf{M} \mathbf{X}_r = 1 \quad (9.96)$$

The modal vector $\mathbf{X}_r$ is then called *orthonormal*. Pre-multiplying both sides of Equation 9.87 by $\mathbf{X}_r^T$ yields

$$\mathbf{X}_r^T \mathbf{K} \mathbf{X}_r = \omega_r^2 \quad (9.97)$$

Example 9.10: Normalization of Modal Vectors

Normalize the modal vectors $\mathbf{X}_1$, $\mathbf{X}_2$, and $\mathbf{X}_3$ obtained in Part (a) of Example 9.9 to a set of orthonormal modal vectors satisfying Equations 9.96 and 9.97. Assume $m = 1$ kg and $k = 1$ N/m.

Solution

Note that regardless of how the modal vectors are normalized, they result in the same mode shape but scaled in magnitude. So, we can write the first modal vector in the form

$$\tilde{\mathbf{X}}_1 = \alpha \mathbf{X}_1$$

where:

$\tilde{\mathbf{X}}_1$ is the modal vector after normalization

α is a non-zero scaling constant

Substituting $\tilde{\mathbf{X}}_1$ into Equation 9.96 gives

$$\tilde{\mathbf{X}}_1^T \mathbf{M} \tilde{\mathbf{X}}_1 = (\alpha \mathbf{X}_1)^T \mathbf{M} (\alpha \mathbf{X}_1) = \alpha^2 \mathbf{X}_1^T \mathbf{M} \mathbf{X}_1$$

which should be equal to 1 if $\tilde{\mathbf{X}}_1$ is orthonormal. Inserting $\mathbf{X}_1 = [0.2549 \ 0.5842 \ 1]^T$ and $\mathbf{M}$ obtained in Example 9.9 yields

$$\alpha^2 \mathbf{X}_1^T \mathbf{M} \mathbf{X}_1 = \alpha^2 \begin{bmatrix} 0.2549 \\ 0.5842 \\ 1 \end{bmatrix}^T \begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix} \begin{bmatrix} 0.2549 \\ 0.5842 \\ 1 \end{bmatrix} = 1.4063 \alpha^2 = 1$$

Thus, $\alpha = 0.8433$ and

$$\tilde{\mathbf{X}}_1 = \alpha \mathbf{X}_1 = \begin{bmatrix} 0.2150 \\ 0.4927 \\ 0.8433 \end{bmatrix}$$

Neglecting round-off error, note that $\tilde{\mathbf{X}}_1$ is the same as the first column in matrix $\mathbf{V}$ obtained in Example 9.9, Part (b). Following the same procedure yields the other two orthonormal modal vectors, $\tilde{\mathbf{X}}_2$ and $\tilde{\mathbf{X}}_3$,

$$\tilde{\mathbf{X}}_2 = 0.6831\mathbf{X}_2 = \begin{bmatrix} 0.5049 \\ 0.6831 \\ -0.5278 \end{bmatrix}, \quad \tilde{\mathbf{X}}_3 = 0.8360\mathbf{X}_3 = \begin{bmatrix} 0.8360 \\ -0.5392 \\ 0.1019 \end{bmatrix}$$

which correspond to the second and third columns in matrix $\mathbf{V}$ with all elements multiplied by -1 . Thus, the MATLAB command `eig` returns a set of orthonormal modal vectors.

The modal vectors $\mathbf{X}_1, \dots, \mathbf{X}_n$ are generally assembled side by side to form the modal matrix

$$\Phi = [\mathbf{X}_1 \cdots \mathbf{X}_n]$$

If the modal vectors are orthonormal, then we have

$$\Phi^T \mathbf{M} \Phi = \begin{bmatrix} \tilde{\mathbf{X}}_1^T \\ \vdots \\ \tilde{\mathbf{X}}_n^T \end{bmatrix} \mathbf{M} \begin{bmatrix} \tilde{\mathbf{X}}_1 & \cdots & \tilde{\mathbf{X}}_n \end{bmatrix} = \begin{bmatrix} \tilde{\mathbf{X}}_1^T \mathbf{M} \tilde{\mathbf{X}}_1 & \cdots & \tilde{\mathbf{X}}_1^T \mathbf{M} \tilde{\mathbf{X}}_n \\ \vdots & \ddots & \vdots \\ \tilde{\mathbf{X}}_n^T \mathbf{M} \tilde{\mathbf{X}}_1 & \cdots & \tilde{\mathbf{X}}_n^T \mathbf{M} \tilde{\mathbf{X}}_n \end{bmatrix} \quad (9.98)$$

Due to the orthogonality of modal vectors as defined in Equation 9.94, all off-diagonal entries are zero. Also, the modal vectors $\tilde{\mathbf{X}}_1, \dots, \tilde{\mathbf{X}}_n$ are orthonormal. Following Equation 9.96, each entry along the diagonal should be 1. Thus,

$$\Phi^T \mathbf{M} \Phi = \mathbf{I}_n \quad (9.99)$$

Similarly, we have

$$\Phi^T \mathbf{K} \Phi = \begin{bmatrix} \tilde{\mathbf{X}}_1^T \\ \vdots \\ \tilde{\mathbf{X}}_n^T \end{bmatrix} \mathbf{K} \begin{bmatrix} \tilde{\mathbf{X}}_1 & \cdots & \tilde{\mathbf{X}}_n \end{bmatrix} = \begin{bmatrix} \tilde{\mathbf{X}}_1^T \mathbf{K} \tilde{\mathbf{X}}_1 & \cdots & \tilde{\mathbf{X}}_1^T \mathbf{K} \tilde{\mathbf{X}}_n \\ \vdots & \ddots & \vdots \\ \tilde{\mathbf{X}}_n^T \mathbf{K} \tilde{\mathbf{X}}_1 & \cdots & \tilde{\mathbf{X}}_n^T \mathbf{K} \tilde{\mathbf{X}}_n \end{bmatrix} \quad (9.100)$$

Following Equations 9.95 and 9.97,

$$\Phi^T \mathbf{K} \Phi = \begin{bmatrix} \omega_1^2 & 0 & 0 \\ 0 & \ddots & 0 \\ 0 & 0 & \omega_n^2 \end{bmatrix} = \Omega \quad (9.101)$$

Equations 9.99 and 9.101 are very useful for decoupling equations of motion and obtaining responses of vibration systems.

9.4.3 Response to Initial Excitations

Now, let us consider the free, undamped vibration problem described by Equation 9.81, in which the initial condition vectors are $\mathbf{x}(0)$ and $\dot{\mathbf{x}}(0)$. Solving the associated eigenvalue problem yields a set of natural frequencies, $\omega_1, \omega_2, \dots$, and ω_n , and a set of orthonormal modal vectors, $\tilde{\mathbf{X}}_1, \tilde{\mathbf{X}}_2, \dots$, and $\tilde{\mathbf{X}}_n$. To decouple the equations of motion, we introduce a coordinate transformation

$$\mathbf{x} = \Phi \mathbf{q} \quad (9.102)$$

where:

$\Phi = [\tilde{\mathbf{X}}_1 \ \tilde{\mathbf{X}}_2 \ \dots \ \tilde{\mathbf{X}}_n]$ is the modal matrix
 $\mathbf{q}$ is the vector of modal coordinates

Substituting Equation 9.102 into Equation 9.81 and pre-multiplying by Φ^T on both sides gives

$$\Phi^T \mathbf{M} \Phi \ddot{\mathbf{q}} + \Phi^T \mathbf{K} \Phi \mathbf{q} = \mathbf{0} \quad (9.103)$$

Due to the orthogonality of modal vectors given by Equations 9.99 and 9.101, Equation 9.103 reduces to

$$\ddot{\mathbf{q}} + \Omega \mathbf{q} = \mathbf{0} \quad (9.104)$$

which can be written as n independent modal equations

$$\ddot{q}_r + \omega_r^2 q_r = 0, \quad r = 1, 2, \dots, n \quad (9.105)$$

Obviously, modal equations are analogous to differential equations of single-degree-of-freedom systems, for which the response was given in [Section 8.3](#). Following the discussion in [Chapter 8](#), the solutions to Equation 9.105 are

$$q_r(t) = q_r(0) \cos(\omega_r t) + \frac{\dot{q}_r(0)}{\omega_r} \sin(\omega_r t), \quad r = 1, 2, \dots, n \quad (9.106)$$

where $q_r(0)$ and $\dot{q}_r(0)$ are the initial displacements and initial velocities of modal coordinates. Note that the prescribed initial conditions are associated with the physical coordinates $\mathbf{x}$. To transform the initial conditions, we use Equation 9.102, which gives $\mathbf{q} = \Phi^{-1} \mathbf{x}$. Because of the property of orthonormal modal vectors, $\Phi^T \mathbf{M} \Phi = \mathbf{I}_n$, we have $\Phi^{-1} = \Phi^T \mathbf{M}$. Thus, the initial modal displacements and velocities are

$$\mathbf{q}(0) = \Phi^T \mathbf{M} \mathbf{x}(0) \quad (9.107)$$

$$\dot{\mathbf{q}}(0) = \Phi^T \mathbf{M} \dot{\mathbf{x}}(0) \quad (9.108)$$

Expanding the modal matrix yields

$$\begin{Bmatrix} q_1(0) \\ \vdots \\ q_n(0) \end{Bmatrix} = \begin{bmatrix} \tilde{\mathbf{X}}_1^T \\ \vdots \\ \tilde{\mathbf{X}}_n^T \end{bmatrix} \mathbf{M} \mathbf{x}(0) = \begin{Bmatrix} \tilde{\mathbf{X}}_1^T \mathbf{M} \mathbf{x}(0) \\ \vdots \\ \tilde{\mathbf{X}}_n^T \mathbf{M} \mathbf{x}(0) \end{Bmatrix} \quad (9.109)$$

$$\begin{Bmatrix} \dot{q}_1(0) \\ \vdots \\ \dot{q}_n(0) \end{Bmatrix} = \begin{bmatrix} \tilde{\mathbf{X}}_1^T \\ \vdots \\ \tilde{\mathbf{X}}_n^T \end{bmatrix} \mathbf{M} \dot{\mathbf{x}}(0) = \begin{Bmatrix} \tilde{\mathbf{X}}_1^T \mathbf{M} \dot{\mathbf{x}}(0) \\ \vdots \\ \tilde{\mathbf{X}}_n^T \mathbf{M} \dot{\mathbf{x}}(0) \end{Bmatrix} \quad (9.110)$$

or

$$q_r(0) = \tilde{\mathbf{X}}_r^T \mathbf{M}\mathbf{x}(0), \quad r = 1, 2, \dots, n \quad (9.111)$$

$$\dot{q}_r(0) = \tilde{\mathbf{X}}_r^T \mathbf{M}\dot{\mathbf{x}}(0), \quad r = 1, 2, \dots, n \quad (9.112)$$

Then, the modal responses are

$$q_r(t) = \tilde{\mathbf{X}}_r^T \mathbf{M}\mathbf{x}(0) \cos(\omega_r t) + \frac{1}{\omega_r} \tilde{\mathbf{X}}_r^T \mathbf{M}\dot{\mathbf{x}}(0) \sin(\omega_r t), \quad r = 1, 2, \dots, n \quad (9.113)$$

Finally, application of Equation 9.102 gives the responses of n -degree-of-freedom systems to initial excitations

$$\mathbf{x}(t) = \begin{bmatrix} \tilde{\mathbf{X}}_1 & \dots & \tilde{\mathbf{X}}_n \end{bmatrix} \begin{Bmatrix} q_1(t) \\ \vdots \\ q_n(t) \end{Bmatrix} = \sum_{r=1}^n \tilde{\mathbf{X}}_r q_r(t) = \sum_{r=1}^n q_r(t) \tilde{\mathbf{X}}_r \quad (9.114)$$

Equation 9.114 represents the so-called expansion theorem, by which the solution $\mathbf{x}(t)$ can be regarded as a superposition of the normal modes $\tilde{\mathbf{X}}_r(t)$ ($r = 1, 2, \dots, n$). The modal responses $q_r(t)$ represent the contributions of the particular configuration $\tilde{\mathbf{X}}_r(t)$ to the total solution.

The approach using the orthogonality properties of the modal matrix to obtain a set of simultaneous independent equations is known as modal analysis. The basic steps to obtain the solutions by using modal analysis are summarized as follows:

Step 1: Solve the eigenvalue problem and obtain the natural frequencies and modal vectors.

Step 2: Normalize the modal vectors to obtain orthonormality.

Step 3: Determine the modal responses and combine them into the response of the original system by using the expansion theorem.

Example 9.11: Response to Initial Excitation by Modal Analysis

Consider the three-degree-of-freedom system discussed in Example 9.9. Assume $m = 1$ kg and $k = 1$ N/m. Using the natural frequencies obtained in Example 9.9 and the orthonormal modal vectors obtained in Example 9.10, determine the response of the system subjected to initial excitations $\mathbf{x}(0) = [0 \ 0 \ 0.01]^T$ and $\dot{\mathbf{x}}(0) = [0 \ 0 \ 0]^T$.

Solution

In Examples 9.9 and 9.10, we obtained the three natural frequencies

$$\omega_1 = 0.6448 \text{ rad/s}, \quad \omega_2 = 1.5147 \text{ rad/s}, \quad \omega_3 = 2.5080 \text{ rad/s}$$

and the three orthonormal modal vectors

$$\tilde{\mathbf{X}}_1 = \begin{bmatrix} 0.2150 \\ 0.4927 \\ 0.8433 \end{bmatrix}, \quad \tilde{\mathbf{X}}_2 = \begin{bmatrix} 0.5049 \\ 0.6831 \\ -0.5278 \end{bmatrix}, \quad \tilde{\mathbf{X}}_3 = \begin{bmatrix} 0.8360 \\ -0.5392 \\ 0.1019 \end{bmatrix}$$

Applying Equation 9.113 gives the modal responses

$$\begin{aligned}
 q_1(t) &= \begin{bmatrix} 0.2150 & 0.4927 & 0.8433 \end{bmatrix} \begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix} \begin{Bmatrix} 0 \\ 0 \\ 0.01 \end{Bmatrix} \cos(0.6448t) = 0.0084 \cos(0.6448t) \\
 q_2(t) &= \begin{bmatrix} 0.5049 & 0.6831 & -0.5278 \end{bmatrix} \begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix} \begin{Bmatrix} 0 \\ 0 \\ 0.01 \end{Bmatrix} \cos(1.5147t) = -0.0053 \cos(1.5147t) \\
 q_3(t) &= \begin{bmatrix} 0.8360 & -0.5392 & 0.1019 \end{bmatrix} \begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix} \begin{Bmatrix} 0 \\ 0 \\ 0.01 \end{Bmatrix} \cos(2.5080t) = 0.0010 \cos(2.5080t)
 \end{aligned}$$

Inserting the modal responses into Equation 9.114, we obtain the response of the system to the given initial excitations as follows.

$$\begin{aligned}
 \mathbf{x}(t) &= 0.0084 \cos(0.6448t) \begin{bmatrix} 0.2150 \\ 0.4927 \\ 0.8433 \end{bmatrix} - 0.0053 \cos(1.5147t) \begin{bmatrix} 0.5049 \\ 0.6831 \\ -0.5278 \end{bmatrix} \\
 &\quad + 0.0010 \cos(2.5080t) \begin{bmatrix} 0.8360 \\ -0.5392 \\ 0.1019 \end{bmatrix} \\
 &= \begin{bmatrix} 0.0018 \\ 0.0041 \\ 0.0071 \end{bmatrix} \cos(0.6448t) + \begin{bmatrix} -0.0027 \\ -0.0036 \\ 0.0028 \end{bmatrix} \cos(1.5147t) + \begin{bmatrix} 0.0084 \\ -0.0054 \\ 0.0010 \end{bmatrix} \cos(2.5080t)
 \end{aligned}$$

9.4.4 Response to Harmonic Excitations

The essence of modal analysis is to determine the response of an n -degree-of-freedom system by decomposing it into n single-degree-of-freedom systems, determining the response of each single-degree-of-freedom system, and combining the individual responses into the response of the original system. The previous subsection showed how one can obtain the free response of an undamped system by using modal analysis, which can also be used to find the response of an undamped system to harmonic excitations. In this case, the equations of motion in matrix form are

$$\mathbf{M}\ddot{\mathbf{x}} + \mathbf{K}\mathbf{x} = \mathbf{f} \quad (9.115)$$

where $\mathbf{f}$ is an $n \times 1$ force vector. Following the basic steps of modal analysis given in [Subsection 9.4.3](#), we first need to solve the eigenvalue problem $(\mathbf{K} - \omega^2\mathbf{M})\mathbf{X} = \mathbf{0}$ and obtain the solutions ω_r^2 and $\mathbf{X}_r$ ($r = 1, 2, \dots, n$). The n modal vectors are then normalized and arranged into the $n \times n$ modal matrix $\Phi = [\tilde{\mathbf{X}}_1 \cdots \tilde{\mathbf{X}}_n]$, which satisfies the orthonormality relations $\Phi^T\mathbf{M}\Phi = \mathbf{I}_n$ and $\Phi^T\mathbf{K}\Phi = \mathbf{\Omega}$. Using the expansion theorem, the solution

to Equation 9.115 is a linear combination of the modal vectors, $\mathbf{x} = \sum_{r=1}^n q_r(t) \widetilde{\mathbf{X}}_r = \Phi \mathbf{q}$. Substituting it into Equation 9.115 and pre-multiplying the result by Φ^T gives

$$\Phi^T \mathbf{M} \Phi \ddot{\mathbf{q}} + \Phi^T \mathbf{K} \Phi \mathbf{q} = \Phi^T \mathbf{f} \quad (9.116)$$

Combining the orthonormality relations, we have

$$\ddot{\mathbf{q}} + \Omega \mathbf{q} = \mathbf{N} \quad (9.117)$$

where $\mathbf{N} = \Phi^T \mathbf{f}$. Note that

$$\Phi^T \mathbf{f} = \begin{bmatrix} \widetilde{\mathbf{X}}_1^T \\ \vdots \\ \widetilde{\mathbf{X}}_n^T \end{bmatrix} \mathbf{f} = \left\{ \begin{bmatrix} \widetilde{\mathbf{X}}_1^T \mathbf{f} \\ \vdots \\ \widetilde{\mathbf{X}}_n^T \mathbf{f} \end{bmatrix} \right\} = \left\{ \begin{bmatrix} N_1 \\ \vdots \\ N_n \end{bmatrix} \right\} \quad (9.118)$$

where the entries $\widetilde{\mathbf{X}}_r^T \mathbf{f}$ ($r = 1, 2, \dots, n$) are the products of the $1 \times n$ row vectors $\widetilde{\mathbf{X}}_r^T$ and the $n \times 1$ column vector $\mathbf{f}$, resulting in scalars N_r , known as modal forces. Equation 9.117 is equivalent to a set of n independent modal equations

$$\ddot{q}_r + \omega_r^2 q_r = N_r, \quad r = 1, 2, \dots, n \quad (9.119)$$

We consider harmonic excitations, without loss of generality, in the form

$$\mathbf{f} = \mathbf{f}_0 \sin(\omega t) \quad (9.120)$$

Then, the modal forces N_r can be written as $\widetilde{\mathbf{X}}_r^T \mathbf{f}_0 \sin(\omega t)$ ($r = 1, 2, \dots, n$). The solutions to Equation 9.119 can be obtained in a manner similar to the response of undamped single-degree-of-freedom systems to harmonic excitations. Applying Equations 9.31 through 9.33 with $\zeta = 0$, we have

$$q_r(t) = \frac{\widetilde{\mathbf{X}}_r^T \mathbf{f}_0}{\omega_r^2 - \omega^2} \sin(\omega t), \quad r = 1, 2, \dots, n \quad (9.121)$$

Thus, the steady-state response of the original system is

$$\mathbf{x} = \sum_{r=1}^n q_r(t) \widetilde{\mathbf{X}}_r = \sum_{r=1}^n \frac{\widetilde{\mathbf{X}}_r^T \mathbf{f}_0 \widetilde{\mathbf{X}}_r}{\omega_r^2 - \omega^2} \sin(\omega t) \quad (9.122)$$

Example 9.12: Response to Harmonic Excitation by Modal Analysis

Consider the three-degree-of-freedom system discussed in Example 9.9. Assume $m = 1$ kg and $k = 1$ N/m. Using the natural frequencies obtained in Example 9.9 and the orthonormal modal vectors obtained in Example 9.10, determine the response of the system if a harmonic force $F(t) = 3\sin(2t)$ is applied to mass 3.

Solution

The force vector can be written as

$$\mathbf{f} = \begin{Bmatrix} 0 \\ 0 \\ F(t) \end{Bmatrix} = \begin{Bmatrix} 0 \\ 0 \\ 3\sin(2t) \end{Bmatrix} = \begin{Bmatrix} 0 \\ 0 \\ 3 \end{Bmatrix} \sin(2t)$$

From Examples 9.9 and 9.10, the three natural frequencies are

$$\omega_1 = 0.6448 \text{ rad/s}, \quad \omega_2 = 1.5147 \text{ rad/s}, \quad \omega_3 = 2.5080 \text{ rad/s}$$

and the three orthonormal modal vectors are

$$\tilde{\mathbf{X}}_1 = \begin{bmatrix} 0.2150 \\ 0.4927 \\ 0.8433 \end{bmatrix}, \quad \tilde{\mathbf{X}}_2 = \begin{bmatrix} 0.5049 \\ 0.6831 \\ -0.5278 \end{bmatrix}, \quad \tilde{\mathbf{X}}_3 = \begin{bmatrix} 0.8360 \\ -0.5392 \\ 0.1019 \end{bmatrix}$$

Equation 9.121 gives the modal responses

$$q_1(t) = \frac{\begin{bmatrix} 0.2150 & 0.4927 & 0.8433 \end{bmatrix} \begin{Bmatrix} 0 \\ 0 \\ 3 \end{Bmatrix}}{0.6448^2 - 2^2} \sin(2t) = -0.7058 \sin(2t)$$

$$q_2(t) = \frac{\begin{bmatrix} 0.5049 & 0.6831 & -0.5278 \end{bmatrix} \begin{Bmatrix} 0 \\ 0 \\ 3 \end{Bmatrix}}{1.5147^2 - 2^2} \sin(2t) = 0.9283 \sin(2t)$$

$$q_3(t) = \frac{\begin{bmatrix} 0.8360 & -0.5392 & 0.1019 \end{bmatrix} \begin{Bmatrix} 0 \\ 0 \\ 3 \end{Bmatrix}}{2.5080^2 - 2^2} \sin(2t) = 0.1335 \sin(2t)$$

Inserting the modal responses into Equation 9.122, we obtain the response of the system to the given harmonic excitation as follows.

$$\begin{aligned} \mathbf{x}(t) &= -0.7058 \sin(2t) \begin{bmatrix} 0.2150 \\ 0.4927 \\ 0.8433 \end{bmatrix} + 0.9283 \sin(2t) \begin{bmatrix} 0.5049 \\ 0.6831 \\ -0.5278 \end{bmatrix} + 0.1335 \sin(2t) \begin{bmatrix} 0.8360 \\ -0.5392 \\ 0.1019 \end{bmatrix} \\ &= \begin{Bmatrix} 0.4286 \\ 0.2144 \\ -1.0716 \end{Bmatrix} \sin(2t) \end{aligned}$$

PROBLEM SET 9.4

1. Consider the two-degree-of-freedom mass–spring system shown in Figure 9.22. Assume that $m_1 = m$, $m_2 = 2m$, and $k_1 = k_2 = k$, where $m = 5$ kg and $k = 2000$ N/m.
 - a. Derive the equations of motion and express them in matrix form.
 - b. Solve the associated eigenvalue problem by hand. Plot the two modes and explain the nature of the mode shapes.
 - c.  Solve the associated eigenvalue problem by using MATLAB.
2. A three-story building can be modeled as a three-degree-of-freedom system, as shown in Figure 9.23, in which the horizontal members are rigid and the columns are massless beams acting as springs. Assume that $m_1 = 1200$ kg, $m_2 = 2400$ kg, $m_3 = 3600$ kg, $k_1 = 500$ kN/m, $k_2 = 1000$ kN/m, and $k_3 = 1500$ kN/m.
 - a. Derive the differential equations for the horizontal motion of the masses.
 - b. Solve the associated eigenvalue problem by hand. Plot the three modes and explain the nature of the mode shapes.
 - c.  Solve the associated eigenvalue problem by using MATLAB.

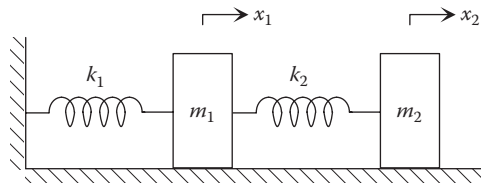


FIGURE 9.22
Problem 1.

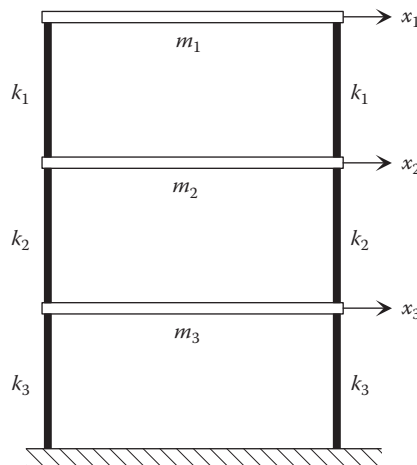


FIGURE 9.23
Problem 2.

3. Consider the two-degree-of-freedom mass–spring system in [Figure 9.22](#). Normalize the modal vectors $\mathbf{X}_1$ and $\mathbf{X}_2$ obtained in Part (b) of Problem 1 to a set of orthonormal modal vectors satisfying Equations 9.96 and 9.97.
4. Consider the three-story building system in [Figure 9.23](#). Normalize the modal vectors $\mathbf{X}_1$, $\mathbf{X}_2$, and $\mathbf{X}_3$ obtained in Part (b) of Problem 2 to a set of orthonormal modal vectors satisfying Equations 9.96 and 9.97.
5. Consider the two-degree-of-freedom mass–spring system in [Figure 9.22](#). Determine the response of the system subjected to initial excitations $\mathbf{x}(0) = [0 \ 0]^T$ and $\dot{\mathbf{x}}(0) = [1 \ 0]^T$ by modal analysis. Note that the natural frequencies were found in Problem 1 and the modal vectors were normalized in Problem 3.
6. Consider the three-story building system in [Figure 9.23](#). Determine the response of the system subjected to initial excitations $\mathbf{x}(0) = [0.01 \ 0 \ 0]^T$ and $\dot{\mathbf{x}}(0) = [0 \ 0 \ 0]^T$ by modal analysis. Note that the natural frequencies were found in Problem 2 and the modal vectors were normalized in Problem 4.
7. Consider the two-degree-of-freedom mass–spring system in [Figure 9.22](#). Find the response of the system if a harmonic force $F(t) = 2 \cos(3t)$ is applied to mass 2. Note that the natural frequencies were found in Problem 1 and the modal vectors were normalized in Problem 3.
8. Consider the three-story building system in [Figure 9.23](#). Determine the response of the system if a harmonic force $F(t) = 15 \sin(0.1t)$ is applied to the top story. Note that the natural frequencies were found in Problem 2 and the modal vectors were normalized in Problem 4.
9.  Solve Problem 5 by using MATLAB.
10.  Solve Problem 6 by using MATLAB.
11.  Solve Problem 7 by using MATLAB.
12.  Solve Problem 8 by using MATLAB.

9.5 Vibration Measurement and Analysis

Models are necessary for designing dynamic systems and understanding system dynamics. For many vibration systems, theoretical models are too difficult to develop. In these situations, we resort to experimental models, which may be obtained using experimental modal techniques; specifically, the method known as frequency response function testing. There are two phases needed to obtain an experimental model of a vibration system. The first phase is the measurement phase, in which the frequency response functions of the system are measured. The second phase is the analysis phase, in which system parameters are estimated from the measured frequency response functions. In this section, we first introduce the measurement methods for acquiring frequency response data and then discuss the parameter estimation methods for extracting system properties.

9.5.1 Vibration Measurement

As presented in [Section 8.4](#), a frequency response function is a transfer function, expressed in the frequency domain as opposed to the s domain. For a vibration system, a frequency

response function describes the system response to an external excitation force as a function of frequency. The response may be displacement, velocity, or acceleration. To experimentally obtain the frequency response function of a vibration system, the input excitation and output response must be measured simultaneously. The excitation and response are both obtained in the frequency domain via fast Fourier transform, and the frequency response function is the ratio of the response to the excitation.

Figure 9.24 shows a diagram of the basic components for frequency response function measurement. An actuator, or so-called exciter, is used to apply a force to the system under test. Sensors, or so-called transducers, are used to measure the force and responses. A data acquisition system is used to acquire and process the signals from the sensors. A computer with analysis software provides measurement functions such as windowing, averaging, and fast Fourier transform computation.

The first step in the measurement process involves selecting an excitation function along with an excitation system. The excitation function is the mathematical signal used for the input. The excitation system is the physical mechanism used to provide the signal. Two of the general categories of excitation functions are steady-state and transient. For example, a sine function is a steady-state signal and an impulse function is a transient signal. Two of the most common excitation mechanisms are a shaker and an impact hammer.

Figure 9.25 shows an electromagnetic shaker, which is one of the most common shakers for vibration testing. With the electromagnetic shaker, a force is generated by an alternating current that drives a magnetic coil. Such a shaker can generate a variety of time-varying forces, such as a sinusoidal force with a constant frequency and a swept sine function with gradually increasing frequency but constant amplitude. The maximum frequency range and the maximum force level depend on the size of the shaker. Smaller shakers have a

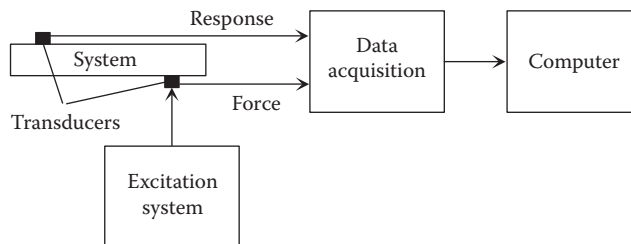


FIGURE 9.24

Basic components for frequency response function measurement.

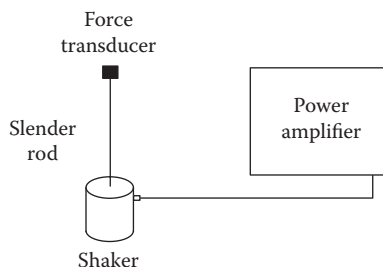
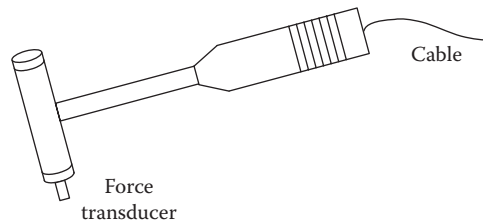


FIGURE 9.25

An electromagnetic shaker with a power amplifier.

**FIGURE 9.26**

An impact hammer for vibration testing.

higher frequency range and a lower force level. When using a shaker for excitation, the shaker should be physically mounted on the system via a force transducer.

An impact hammer is another common excitation mechanism, which is used to apply impulse. As shown in [Figure 9.26](#), the impact hammer has a transducer at the tip for measuring the impact force. If the system being tested is struck by the hammer quickly, the resulting excitation force resembles an impulse. Because the impact hammer does not have to be attached to the system under test, this technique is relatively easy to implement, but it is difficult to obtain consistent results. The frequency content and the pulse duration depend on the material of the tip. The harder the tip, the shorter the pulse duration included, and thus, the higher the frequency content measured.

The second step in the measurement process is to select the transducers for sensing force and response. The piezoelectric type is the most widely used for vibration testing. The piezoelectric transducer is an electromechanical sensor that generates an electrical output when subjected to vibration. The response measured is usually acceleration. This particular type of response transducer is called an accelerometer. Sometimes, the system is too small, or the working environment is too hot, to attach an accelerometer to the system under consideration. In those situations, a laser Doppler vibrometer can be used to make noncontact vibration measurements.

With the selected excitation mechanism and transducers, the frequency response function can be obtained in several different ways. For example, if a shaker is used, then harmonic excitation is applied to the system under test and the resulting harmonic response is measured. This type of test is referred to as sine wave testing. The frequency range is covered either by stepping from one frequency to the next or by slowly sweeping the frequency continuously. In both cases, the measurement time should be long enough to allow steady-state conditions to be attained. If an impact hammer is used, then impulsive excitation is applied to the system under test and the resulting transient response is measured.

9.5.2 Identification of System Parameters

After having acquired frequency response data via vibration testing, the next major step is to identify the system parameters, more specifically, modal parameters. The basic information that can be determined from frequency response functions includes natural frequencies, the damping ratio associated with each mode, and mode shapes. The discussions in this subsection are only limited to the identification of natural frequencies and damping ratios.

To obtain an accurate estimation, it is important to understand the relationships between frequency response functions and their individual modal parameters. The basics

of a single-degree-of-freedom dynamic system form the basis for parameter estimation techniques. As discussed in [Section 8.4](#), the frequency response function is a complex quantity, which can be presented in terms of magnitude and phase versus frequency. Another method of presenting the frequency response data is to plot the real part and the imaginary part versus frequency. Denote the Fourier transforms of displacement response and excitation force as $X(j\omega)$ and $F(j\omega)$, respectively. The expression of the frequency response function $X(j\omega)/F(j\omega)$ was given in [Section 8.4](#). [Figure 9.27](#) shows the frequency response presented in different forms, where the real and imaginary parts are

$$\operatorname{Re}\left(\frac{X(j\omega)}{F(j\omega)}\right) = \frac{k - m\omega^2}{\sqrt{(k - m\omega^2)^2 + (b\omega)^2}} = \frac{1 - (\omega/\omega_n)^2}{\sqrt{\left[1 - (\omega/\omega_n)^2\right]^2 + (2\zeta\omega/\omega_n)^2}} \quad (9.123)$$

$$\operatorname{Im}\left(\frac{X(j\omega)}{F(j\omega)}\right) = \frac{-b\omega}{\sqrt{(k - m\omega^2)^2 + (b\omega)^2}} = \frac{-2\zeta\omega/\omega_n}{\sqrt{\left[1 - (\omega/\omega_n)^2\right]^2 + (2\zeta\omega/\omega_n)^2}} \quad (9.124)$$

Assume that the system is lightly damped. As shown in [Figure 9.27](#), when the excitation frequency approaches the natural frequency, $\omega = \omega_n$, the system resonates, and the magnitude of the frequency response function reaches its maximum. This conclusion was proved in [Section 8.4](#). Also, at resonance, the real part is equal to zero, and the imaginary part reaches a peak. The former observation can be easily proved by setting $\omega = \omega_n$ in Equation 9.123, whereas the latter will be left as an exercise for the reader. It was discussed earlier in [Section 9.2](#) that the damping ratio ζ can be estimated by the half-power bandwidth method.

In reality, most dynamic systems cannot be simplified as ideal single-degree-of-freedom systems. As presented in [Section 9.4](#), an n -degree-of-freedom system has n modes. For systems with lightly damped and well-separated modes, as shown in [Figure 9.28a](#), the natural frequencies and damping ratios can be estimated using the single-mode method. The basic assumption for single-mode approximation is that in the vicinity of a resonance, the response is due primarily to that single mode. Just like a single-degree-of-freedom

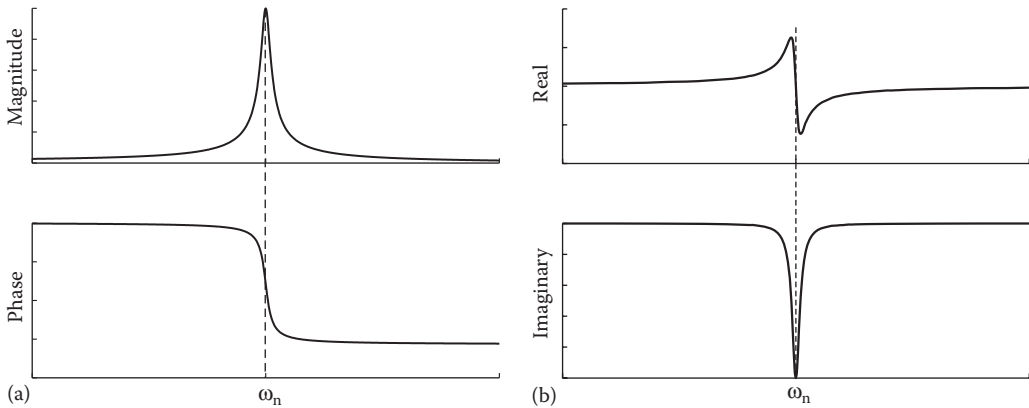
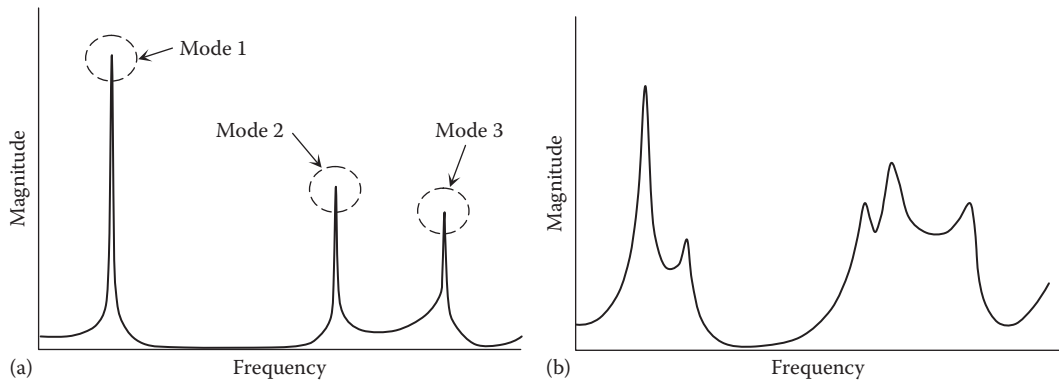


FIGURE 9.27

Frequency response of a single-degree-of-freedom system represented in terms of (a) magnitude and phase and (b) real and imaginary.

**FIGURE 9.28**

Frequency response of a multi-degree-of-freedom system with (a) light damping and (b) heavy damping.

system, the natural frequency associated with that single mode can be estimated from the frequency response data by observing the frequency at which any of the following trends occur:

- The magnitude of the frequency response is a maximum.
- The real part of the frequency response is zero.
- The imaginary part of the frequency response is a maximum or minimum.

The damping ratio associated with that single mode can be estimated using the half-power bandwidth method. For systems with heavily damped and closely spaced modes, as shown in [Figure 9.28b](#), the adjacent modes can affect each other significantly. In general, it will be necessary to implement a multiple-mode method to more accurately identify the modal parameters of these types of systems.

PROBLEM SET 9.5

1. For a single-degree-of-freedom system, denote the Fourier transforms of the displacement response and excitation force as $X(j\omega)$ and $F(j\omega)$, respectively. The expression of the imaginary part of the frequency response function $X(j\omega)/F(j\omega)$ is given by Equation 9.124. Prove that the imaginary part reaches a peak at resonance.
2.  Accelerations are often measured in vibration testing. For a single-degree-of-freedom system, denote the Fourier transforms of the acceleration response and excitation force as $A(j\omega)$ and $F(j\omega)$, respectively. The expression of the frequency response function $A(j\omega)/F(j\omega)$ is given by

$$\frac{A(j\omega)}{F(j\omega)} = \frac{-\omega^2}{k - m\omega^2 + jb\omega}$$

Using MATLAB, write an m-file to plot the magnitude, phase, real part, and imaginary part of the frequency response versus ω/ω_n . Assume that $m = 60$ kg, $b = 25$ N·s/m, and $k = 2000$ N/m.

3. Rods, beams, plates, and so on are continuous systems, which have an infinite number of degrees of freedom and an infinite number of modes. For simplicity, assume that a cantilever beam is approximated as a single-degree-of-freedom mass–damper–spring system, for which the natural frequency is close to the first mode of the beam. The parameters of the cantilever beam are length $L = 0.5$ m, width $b = 0.025$ m, thickness $h = 0.005$ m, density $\rho = 7850$ kg/m³, and Young's modulus $E = 210 \times 10^9$ N/m².
 - a. It is known that the equivalent mass for the beam is $m_{eq} = m/3$, where m is the actual mass of the beam. Determine the equivalent stiffness k_{eq} for the beam. Calculate the natural frequency of the equivalent single-degree-of-freedom system.
 - b. Figure 9.29 is the measured frequency response of the cantilever beam for the first mode. Determine the natural frequency and the damping ratio based on the given information in the plot.
 - c. Compare the frequencies obtained in Parts (a) and (b). What is the error if the cantilever beam is approximated as a single-degree-of-freedom system?
4. Figure 9.30 shows the magnitude of an experimentally determined frequency response. Estimate the number of degrees of freedom of the system and its natural frequencies.

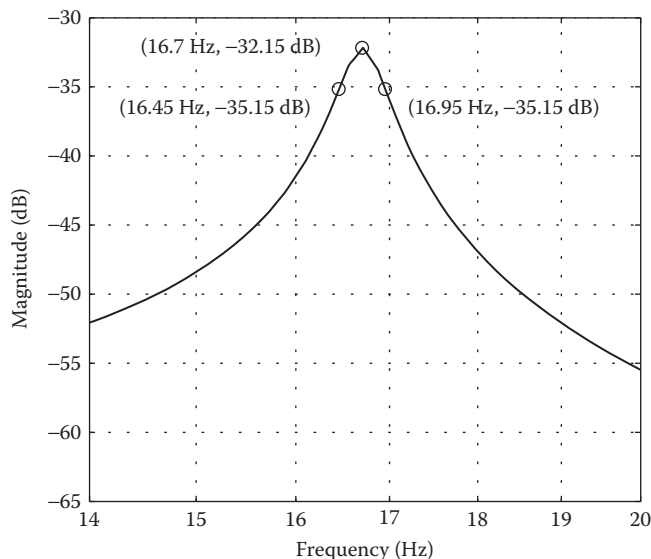


FIGURE 9.29
Problem 3.

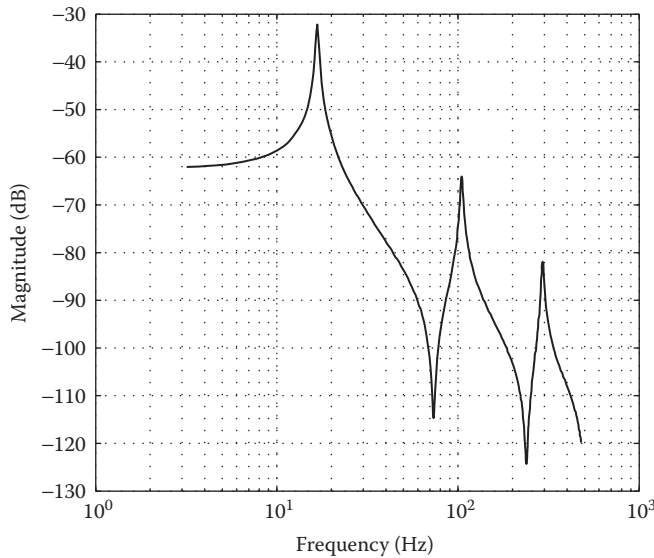


FIGURE 9.30
Problem 4.

9.6 Summary

This chapter presented an introduction to vibrations. Two different types of excitations cause a system to vibrate: initial excitation and external excitation. The vibration of a system caused by non-zero initial displacements or initial velocities (or both) is known as free vibration. The vibration of a system caused by externally applied forces is known as forced vibration.

The nature of system response to initial or external excitations depends on the system's damping, which is a very complex phenomenon. Logarithmic decrement and half-power bandwidth are two commonly used methods for measuring the damping of a vibration system. These two damping identification methods are only valid for the viscous damping model, which is widely used in vibration.

In the logarithmic decrement method, the free response to initial excitations is measured. The damping ratio ζ is estimated using the displacements of two consecutive peaks, x_1 and x_2 . The value of the damping ratio ζ is given by

$$\zeta = \frac{\delta}{\sqrt{(2\pi)^2 + \delta^2}}$$

where δ is called the logarithmic decrement and $\delta = \ln(x_1/x_2)$. For a more accurate estimation, the damping ratio ζ can be determined by measuring the displacements of two peaks separated by a number of periods instead of two consecutive peaks.

In the half-power bandwidth method, the frequency response of the system is measured. The damping ratio ζ is estimated using a peak in the magnitude curve of the frequency

response and two half-power points, which are 3 dB down from the peak. The value of the damping ratio ζ is given by

$$\zeta \approx \frac{\omega_2 - \omega_1}{2\omega_n}$$

where ω_1 and ω_2 are the frequencies at the two half-power points. For light damping, the natural frequency ω_n corresponds to the frequency at the peak or is approximated as $(\omega_1 + \omega_2)/2$.

Frequency response is a very important concept in vibration. For harmonic excitations, more information on the steady-state response can be extracted using the frequency domain technique than the time domain technique. As presented in [Chapter 8](#), the steady-state response of a single-degree-of-freedom mass–damper–spring system to harmonic excitation, for example, $f(t) = F_0 \sin(\omega t)$, is still harmonic with a frequency that is the same as the excitation frequency. Denote the frequency response function of the system as $G(j\omega)$; then, the steady-state response is $x(t) = X \sin(\omega t + \phi)$, where $X = F_0 |G(j\omega)|$ and $\phi = \angle G(j\omega)$. The dimensionless ratio of the dynamic amplitude X and the static deflection x_{st} is given by

$$\frac{X}{x_{st}} = \frac{1}{\sqrt{\left[1 - (\omega/\omega_n)^2\right]^2 + \left[2\zeta(\omega/\omega_n)\right]^2}}$$

Rotating eccentric masses and harmonically moving supports are two common harmonic excitation sources in engineering applications. The steady-state responses to these two excitations can be obtained using the same pattern as in the general case. The results are summarized in [Table 9.1](#). The magnitude of the dimensionless ratios

TABLE 9.1

Summarized Results for Rotating Unbalance and Harmonic Base Excitation

Rotating Unbalance	Harmonic Base Excitation
$M\ddot{x} + b\dot{x} + kx = m\omega^2 \sin(\omega t)$	$\ddot{x}(t) + 2\zeta\omega_n\dot{x}(t) + \omega_n^2 x(t) = 2\zeta\omega_n\dot{z}(t) + \omega_n^2 z(t)$
where the effect of a rotating unbalance mass is to exert a harmonic force $m\omega^2 \sin(\omega t)$	where the motion of the base is harmonic, for example, $z(t) = Z_0 \sin(\omega t)$
$G(j\omega) = \frac{1/k}{1 - (\omega/\omega_n)^2 + j2\zeta\omega/\omega_n}$	$G(j\omega) = \frac{1 + j2\zeta\omega/\omega_n}{1 - (\omega/\omega_n)^2 + j2\zeta\omega/\omega_n}$
$x(t) = X \sin(\omega t + \phi)$	$x(t) = X \sin(\omega t + \phi)$
where	where
$X = \frac{m\omega^2/k}{\sqrt{\left[1 - (\omega/\omega_n)^2\right]^2 + \left[2\zeta(\omega/\omega_n)\right]^2}}$	$X = Z_0 \frac{\sqrt{1 + (2\zeta\omega/\omega_n)^2}}{\sqrt{\left[1 - (\omega/\omega_n)^2\right]^2 + (2\zeta\omega/\omega_n)^2}}$
$\phi = -\tan^{-1} \frac{2\zeta(\omega/\omega_n)}{1 - (\omega/\omega_n)^2}$	$\phi = -\tan^{-1} \frac{2\zeta(\omega/\omega_n)^3}{1 - (\omega/\omega_n)^2 + (2\zeta\omega/\omega_n)^2}$
$\frac{MX}{me} = \frac{(\omega/\omega_n)^2}{\sqrt{\left[1 - (\omega/\omega_n)^2\right]^2 + \left[2\zeta(\omega/\omega_n)\right]^2}}$	$\frac{X}{Z_0} = \frac{\sqrt{1 + (2\zeta\omega/\omega_n)^2}}{\sqrt{\left[1 - (\omega/\omega_n)^2\right]^2 + (2\zeta\omega/\omega_n)^2}}$

MX/me or X/Z_0 versus the driving frequency can be plotted using MATLAB. These plots provide significant information for harmonic responses.

Vibrations are undesirable in many systems. The reduction of vibration can be achieved through vibration isolators or vibration absorbers. A vibration isolation system attempts either to protect delicate equipment from vibration transmitted to it from its support system or to prevent the vibratory force generated by a machine from being transmitted to its surroundings. The essence of these two objectives is the same. The concept of displacement transmissibility X/Z_0 can be used for displacement isolation design, whereas force transmissibility F_T/F_0 can be used for force isolation design, where $X/Z_0 = F_T/F_0$.

To prevent harmonic resonance for a single-degree-of-freedom system, it is not always possible to prevent the natural frequency from being close to the driving frequency by changing either the mass or the spring stiffness. To address this issue, a vibration absorber consisting of a second mass and spring can be added to the system and protect the original single-degree-of-freedom system from harmonic excitation. The vibration of the original mass can be reduced to zero, provided that the natural frequency of the absorber is the same as the driving frequency. One disadvantage of the vibration absorber is that two new resonant frequencies are created.

For multi-degree-of-freedom systems, more advanced mathematical tools are required to solve the equations of motion due to coordinate coupling. Modal analysis is an approach that utilizes the orthogonality of modal vectors to reduce the equations of motion to a set of independent second-order differential equations in modal coordinates. The basic steps in obtaining the response of an n -degree-of-freedom undamped system to initial excitations or external forces by using modal analysis are summarized as follows.

Step 1: Solve the eigenvalue problem associated with the mass and stiffness matrices, that is,

$$(\mathbf{K} - \omega^2 \mathbf{M})\mathbf{X} = \mathbf{0}$$

and obtain the natural frequencies, which are the square roots of eigenvalues $\omega_1^2, \omega_2^2, \dots$, and ω_n^2 , and the modal vectors, which are eigenvectors $\mathbf{X}_1, \mathbf{X}_2, \dots$, and $\mathbf{X}_n$. The natural frequencies are usually arranged in increasing order of magnitude, that is, $\omega_1 \leq \omega_2 \leq \dots \leq \omega_n$. The modal vectors represent the shape of the normal modes physically.

Step 2: Normalize each modal vector to satisfy the relations $\tilde{\mathbf{X}}_r^T \mathbf{M} \tilde{\mathbf{X}}_r = 1$ and $\tilde{\mathbf{X}}_r^T \mathbf{K} \tilde{\mathbf{X}}_r = \omega_r^2$, where $r = 1, 2, \dots, n$. Then, the orthonormal modal matrix $\Phi = [\tilde{\mathbf{X}}_1 \dots \tilde{\mathbf{X}}_n]$ satisfies

$$\Phi^T \mathbf{M} \Phi = \mathbf{I}_n, \quad \Phi^T \mathbf{K} \Phi = \Omega$$

where Ω is a diagonal matrix of the squares of the natural frequencies. Introducing the modal coordinate vector $\mathbf{q}$ and the coordinate transformation $\mathbf{x} = \Phi \mathbf{q}$, we can decouple the original equations as n independent modal equations.

Step 3: Determine the modal responses and combine them into the response of the original system by using the expansion theorem. Table 9.2 lists the responses to initial excitations or external harmonic excitations.

TABLE 9.2

Responses to Initial Excitations or External Harmonic Excitations Using Modal Analysis

	Initial Excitations	External Harmonic Excitations
Equations of motion	$\mathbf{M}\ddot{\mathbf{x}} + \mathbf{K}\mathbf{x} = \mathbf{0}$	$\mathbf{M}\ddot{\mathbf{x}} + \mathbf{K}\mathbf{x} = \mathbf{f}_0 \sin(\omega t)$
Modal equations	$\ddot{q}_r + \omega_r^2 q_r = 0, \quad r = 1, 2, \dots, n$ where the modal initial conditions are $q_r(0) = \tilde{\mathbf{X}}_r^T \mathbf{M}\mathbf{x}(0)$, $\dot{q}_r(0) = \tilde{\mathbf{X}}_r^T \mathbf{M}\dot{\mathbf{x}}(0)$	$\ddot{q}_r + \omega_r^2 q_r = N_r, \quad r = 1, 2, \dots, n$ where the modal forces are $N_r = \tilde{\mathbf{X}}_r^T \mathbf{f}_0 \sin(\omega t)$
Modal responses	$q_r(t) = q_r(0) \cos(\omega_r t) + \frac{\dot{q}_r(0)}{\omega_r} \sin(\omega_r t),$	$q_r(t) = \frac{\tilde{\mathbf{X}}_r^T \mathbf{f}_0}{\omega_r^2 - \omega^2} \sin(\omega t)$
System responses	$\mathbf{x}(t) = \sum_{r=1}^n q_r(t) \tilde{\mathbf{X}}_r$	$\mathbf{x}(t) = \sum_{r=1}^n q_r(t) \tilde{\mathbf{X}}_r$

For many vibration systems, theoretical models are too difficult to develop. In these cases, experimental modal techniques can be used to obtain experimental models. There are two phases involved in obtaining an experimental model of a vibration system. The first phase is the measurement phase, in which the frequency response functions of the system are measured. Two of the most common excitation mechanisms are the shaker and the impact hammer. The second phase is the analysis phase, in which the system parameters are estimated from the measured frequency response functions. The basic information that can be determined from the frequency response functions includes natural frequencies, the damping ratio associated with each mode, and mode shapes.

Review Problems

- Consider the single-degree-of-freedom system shown in [Figure 5.42](#).
 - Determine the undamped natural frequency ω_n , the damping ratio ζ , and the damped natural frequency ω_d .
 - Assume $f(t) = 0$. Find the free vibration response of the system subjected to the initial conditions $x(0) = 0$ m and $\dot{x}(0) = 1$ m/s.
 -  Write a MATLAB m-file to plot the system's response obtained in Part (b).
 -  Construct a Simulink® block diagram based on the differential equation of motion of the system and find the free vibration response.
 -  Build a Simscape® model of the physical system and find the free vibration response.
- Consider the single-degree-of-freedom system shown in [Figure 5.42](#).
 - Assume $f(t) = 500\sin(60t)$ and initial conditions $x(0) = 0$ and $\dot{x}(0) = 0$. Determine the dynamic amplitude X and the static deflection x_{st} of the system. Find the forced vibration response of the system.

- b.  Write a MATLAB m-file to plot the system's response obtained in Part (a).
 - c.  Construct a Simulink block diagram based on the differential equation of motion of the system and find the forced vibration response.
 - d.  Build a Simscape model of the physical system and find the forced vibration response.
 - e.  Assume that the driving frequency varies from 0 to 100 rad/s. Write a MATLAB m-file to plot the dimensionless ratio X/x_{st} versus the frequency ratio ω/ω_n .
3. A machine can be considered as a rigid mass with a rotating unbalanced mass. To reduce the vibration, the machine is mounted on a support system with a stiffness of 10 kN/m. Assume that the total mass is $M = 10$ kg and the unbalanced mass is $m = 0.5$ kg. Determine the range of the damping ratio of the support system so that the vibration amplitude will not exceed 5% of the rotating mass's eccentricity when the machine operates at a speed of 300 rpm.
 4. As shown in [Figure 9.31](#), a machine of a mass $M = 150$ kg is mounted on a fixed-fixed steel beam with negligible mass. The parameters of the beam are given as follows: Young's modulus of the beam $E = 210$ GPa, width $b = 0.75$ m, and thickness $h = 3$ cm. The rotating unbalance in the machine is $me = 0.015$ kg m. The machine runs at a speed of 2400 rpm and the maximum allowable displacement is 4 mm. Determine if it is ok to design the length of the beam as 2 m. Assume damping to be negligible.
 5. An 80-kg machine is placed on a floor that vibrates with a frequency of 50 Hz. The maximum acceleration of the floor is 12.5 cm/s². A vibration isolator consisting of four parallel-connected springs is designed to protect the machine from the vibration of the floor. Assume that the damping ratio of the isolator is 0.1 and the maximum allowable acceleration is 2.5 cm/s². Determine the stiffness of each spring.
 6. Many vibration-measuring instruments consist of a case containing a mass-damper-spring system, as shown in [Figure 9.32](#). The displacement of the mass relative to the case is measured electrically. Denote the displacement of the mass, the displacement of the case, and the displacement of the mass relative to the case as $x(t)$, $z(t)$, and $y(t)$, respectively, where $y(t) = x(t) - z(t)$. Assume harmonic excitation, $z(t) = Z_0 \sin(\omega t)$.

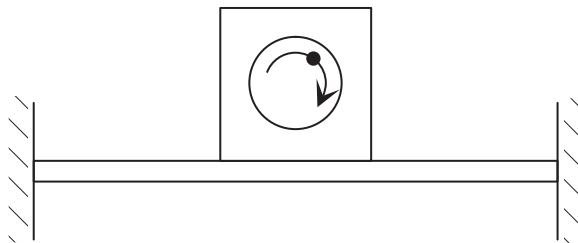
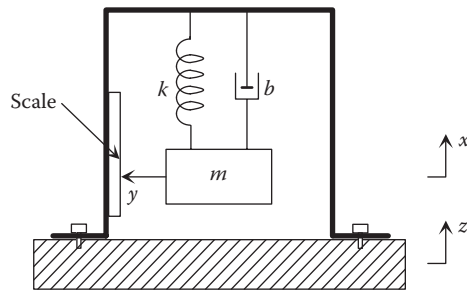


FIGURE 9.31
Problem 4.

**FIGURE 9.32**

Problem 6.

- a. Show that the amplitude Y of $y(t)$ is given by

$$Y = \frac{(\omega/\omega_n)^2}{\sqrt{[1 - (\omega/\omega_n)^2]^2 + (2\zeta\omega/\omega_n)^2}} Z_0$$

- b.  Using MATLAB, write an m-file to plot the frequency response Y/Z_0 versus ω/ω_n for different values of damping ratio: $\zeta = 0.1, 0.25, 0.5, 0.7$, and 1.0 .
7. A 110-kg rotating machine is mounted on a vibration isolator consisting of four parallel-connected springs. The machine operates at a speed of 3000 rpm. Determine the stiffness of each spring so that the force transmitted to the ground is reduced by 90%. Assume that all the springs are identical and damping is negligible.
8. Consider the single-degree-of-freedom system shown in Figure 9.15a, where $m = 10$ kg. When the mass is in equilibrium, the static deformation of the spring is 2 mm. When the mass is allowed to vibrate freely, the maximum displacement amplitude during the fourth cycle is 20% of the first. Assume that a harmonic excitation force is applied to the system.
- Determine the minimum allowable driving frequency if the force transmitted to the ground is less than the excitation force.
 - If the allowable force transmissibility is 10%, determine the stiffness of the spring.
9. Consider the vibration absorber shown in Figure 9.16. It is known that two new resonant frequencies in the neighborhood of the excitation frequency are created. Denote the two new natural frequencies as ω_{n1} and ω_{n2} . Assume $\nu = 1$. Determine ω_{n1}/ω_1 and ω_{n2}/ω_1 for the following cases: $\mu = 0.05, 0.1, 0.15, 0.2$, and 0.25 .
10. A 75-kg table saw is driven by a motor that runs at a constant speed of 1000 rpm and produces a 150-N force. Assume that the stiffness provided by the table legs is 750 kN/m and the damping is negligible.
- Determine the dynamic amplitude of the table.
 - Design a vibration absorber to reduce the table oscillation to zero. Assume that the maximum allowable displacement of the absorber is 2 mm. What is the value of the mass ratio μ ?

11. Consider the linearized model of the double pendulum system in Example 5.16. Assume $m = 1$ kg and $L = 1$ m.
 - a. Use the modal analysis approach to determine the response of the system to the initial excitations $\theta_1(0) = 0.05$ rad, $\theta_2(0) = 0.1$ rad, $\dot{\theta}_1(0) = 0.1$ rad/s, and $\dot{\theta}_2(0) = 0.2$ rad/s.
 - b.  Write a MATLAB m-file to plot the response of the system.
 - c.  Construct a Simulink block diagram based on the linearized model and find the response of the system.
12. Consider the two-degree-of-freedom mass–spring system in Figure 5.118.
 - a. Use the modal analysis approach to determine the response of the system if a harmonic force $f = 40\sin(7\pi t)$ N is applied to mass 1.
 - b.  Write a MATLAB m-file to plot the response of the system.
13. Vibration tests are conducted to estimate parameters of a single-degree-of-freedom mass–damper–spring system.
 - a. The frequency response of the system is shown in Figure 9.33. Determine the system's parameters including the mass m , the stiffness k , and the damping b .
 - b. The free vibration response of the system is given in Figure 9.34, which shows the recorded displacement response of the first three cycles. Using the value of the mass found in Part (a), determine the stiffness k and the damping b .
14. Case study

Consider a small airplane shown in Figure 9.35. Assume that it can be approximated as a four-degree-of-freedom system. The values of the physical parameters are: $m_1 = m_3 = 250$ kg, $m_2 = 1000$ kg, $m_4 = 75$ kg, $k_1 = k_2 = 600$ kN/m (wings), $k_3 = 150$ kN/m (landing-gear system), and $k_4 = 350$ kN/m (tires). In addition,

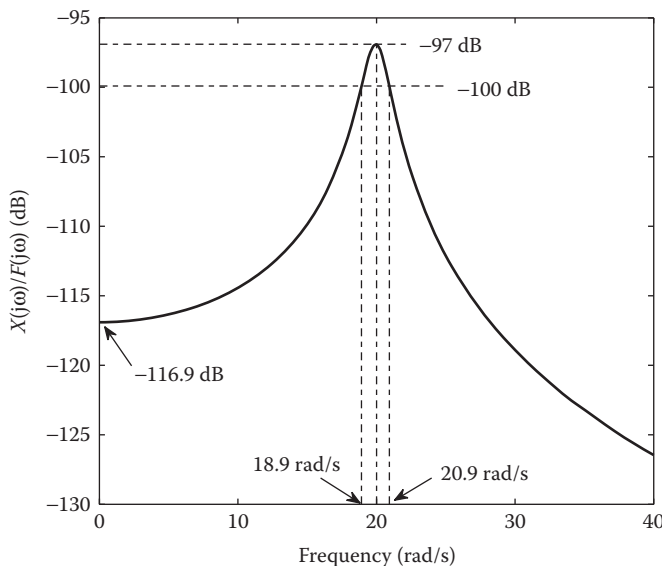


FIGURE 9.33
Problem 13(a).

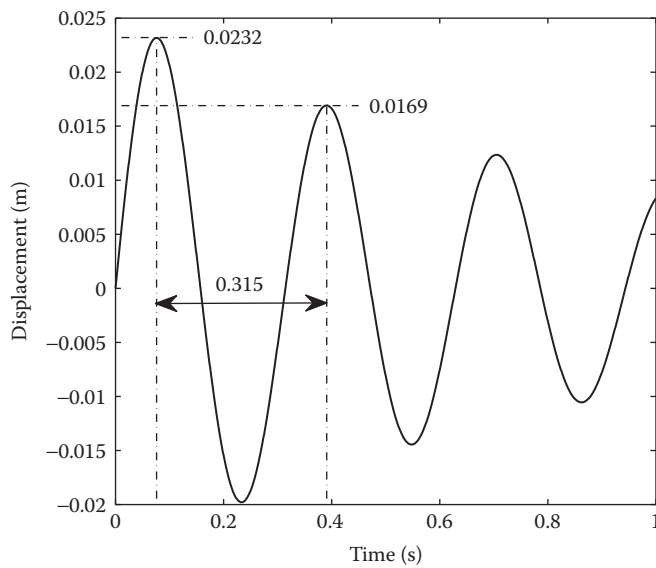


FIGURE 9.34
Problem 13(b).

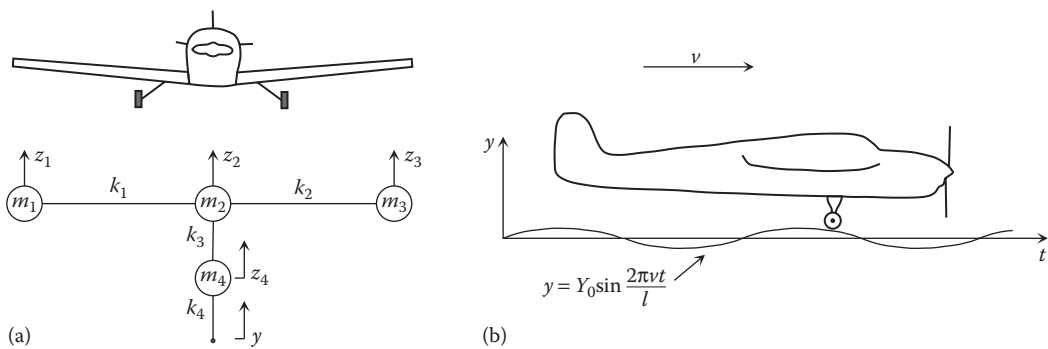


FIGURE 9.35
Problem 14. (a) A small airplane model and (b) touchdown.

it is assumed that the damping coefficient for each mode is 0.05. It is desired to investigate the landing dynamics of this airplane. This may be done by examining both the response of the system to a set of initial conditions and the response to a harmonic input.

- Derive the differential equations of motion for the vertical motion of the airplane.
- ✦ Determine the natural frequencies and mode shapes. Discuss the mode shapes as descriptively as possible.
- ✦ Model the initial touchdown event with the following set of conditions: $\mathbf{z}(0) = [0 \ 0 \ 0 \ 0]^T$ m and $\dot{\mathbf{z}}(0) = [-0.5 \ -0.5 \ -0.5 \ -0.5]^T$ m/s. Plot the motion of each of the masses for the 2 seconds following touchdown.

- d.  The airplane has a normal landing speed of 120 km/h. The runway has a contour that can be modeled by a sine wave with a wavelength of 30 m and amplitude Y_0 of 0.15 m, as shown. Assume that the plane lands at 120 km/h and that it remains in contact with the runway after touchdown so that steady state conditions are established. The input motion to the tires is given by

$$y(t) = Y_0 \sin\left(\frac{2\pi vt}{l}\right)$$

Using the modal analysis approach, solve the forced vibration problem for the steady-state amplitudes of response relative to the runway. Discuss your results.

10

Introduction to Feedback Control Systems

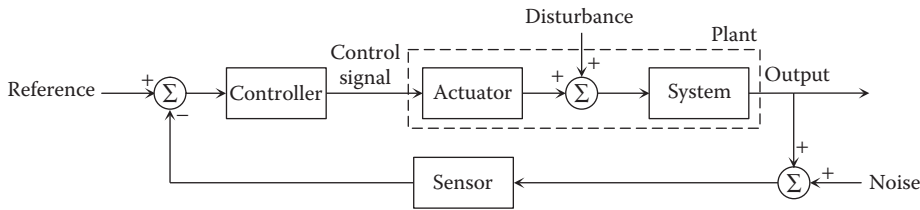
Control deals with the modeling of a variety of dynamic systems and the design of controllers that will ensure that these systems perform in a desired manner. In [Chapters 5 through 7](#), we discussed how to derive a mathematical model for a dynamic system, which can be mechanical, electrical, electromechanical, fluid, or thermal. In this chapter, we focus on how to design a controller for a dynamic system based on its mathematical model. Basic concepts such as feedback, open-loop control, closed-loop control, and basic terminologies in control are introduced in [Section 10.1](#). In general, there are two main reasons why control is needed: one is to maintain the system stability and the other is to improve the system performance. [Section 10.2](#) covers how to determine the stability, how to define the performance in either time domain or frequency domain, and how to identify the model of a system. [Section 10.3](#) discusses the advantages of feedback control, which is utilized in most cases. Following the overview of feedback in [Section 10.3](#), the classical structure of proportional, integral, and derivative control is introduced in [Section 10.4](#). Three different control design methods based on root locus, Bode plot, and state variable feedback are presented in [Sections 10.5 through 10.7](#), respectively. The chapter concludes with controller design and implementation using MATLAB®, Simulink®, and Simscape™ computer tools.

10.1 Basic Concepts and Terminologies

Control is the process of manipulating, manually or automatically, the input of a dynamic system, so that the system output will behave as desired. If the output signal is measured and fed back for use in computing the input signal, the system is called feedback control. A familiar example is the cruise control of an automobile. To maintain a constant vehicle speed set by the driver, the actual speed of the vehicle is measured by the speedometer and fed back to the controller, which adjusts the engine's throttle position. The engine torque is then changed accordingly, which influences the vehicle's actual speed.

To analyze and design a feedback control system, a block diagram is usually drawn to show the major components and their interconnections in graphical form. [Figure 10.1](#) is a general block diagram of an elementary feedback control system. The essential components of this feedback control include a system we want to control, a controller we need to design, an actuator used to drive the controlled system, and a sensor used to measure the system output. The connecting lines in the block diagram carry signals. As shown in [Figure 10.1](#), the important signals in this feedback control system include the output, the control signal, and the reference.

In the example of the automobile's cruise control, the controlled system is the auto-body, whose output is the speed. The speedometer, which acts as a sensor, measures the vehicle speed. The measured speed is fed back and compared with the desired speed, which is the

**FIGURE 10.1**

Block diagram of an elementary feedback control system.

reference signal. Based on the error between the measured and the reference signals, the controller computes the control signal, which is the engine's throttle position in our case. The engine is the actuator, and the torque provided by the engine is applied to the auto-body, which influences the vehicle speed.

Every control system must have these four essential components. Generally, the controlled system and the actuator are intimately connected, and they are combined as one component called the plant. There are two other signals shown in Figure 10.1, disturbance and sensor noise. Both of them are undesired system inputs that adversely affect the performance of a system.

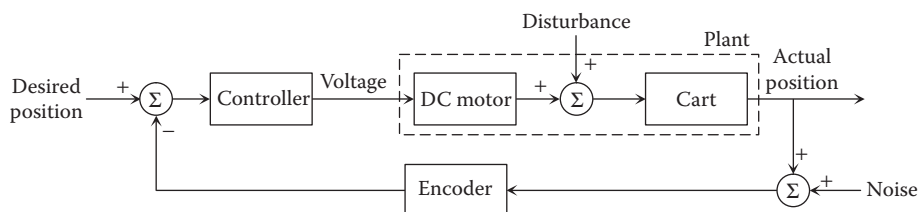
Example 10.1: Block Diagram of a Feedback Control System

Consider the electromechanical system described in Problem 3 of Problem Set 6.4. It consists of a cart that moves along a linear track and a DC motor that drives the cart. An encoder is included to measure the position of the cart. Assume that a controller is designed to control the position of the cart. Draw a block diagram for this feedback control system. Clearly label essential components and signals.

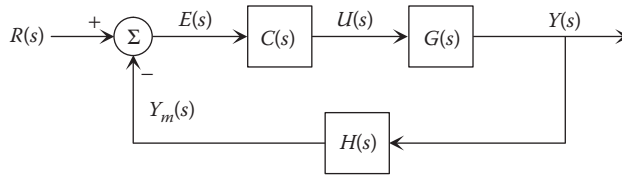
Solution

Note that the essential components of this feedback control are the cart (the controlled system), the controller, the DC motor (the actuator), and the encoder (the sensor). The corresponding block diagram is shown in Figure 10.2. The actual position of the cart is the output, the desired position is the reference, and the voltage applied to the DC motor is the control signal.

Transfer functions are usually used to represent the mathematical model of each block in a block diagram. The input and output signals of each block are also expressed in the Laplace domain, and they are denoted by capital letters. If the disturbance and noise signals are negligible, then the general block diagram of a feedback control system given in Figure 10.1 can be redrawn as shown in Figure 10.3, in which $G(s)$ represents the dynamics

**FIGURE 10.2**

Block diagram of a cart position control system.

**FIGURE 10.3**

General block diagram of a feedback control system, with transfer function representation.

of the plant, $C(s)$ is the controller, $H(s)$ is the sensor, $U(s)$ is the control signal, $Y(s)$ is the actual system output, $Y_m(s)$ is the measured output, and $R(s)$ is the reference. The difference between the reference and the feedback is defined as the error signal,

$$E(s) = R(s) - Y_m(s) \quad (10.1)$$

For an ideal sensor, its output is exactly the same as its input, that is, $H(s) = 1$. Therefore, we have $Y_m(s) = Y(s)$ and

$$E(s) = R(s) - Y(s) \quad (10.2)$$

A control system with feedback is also called closed-loop control. If the feedback is subtracted, it is called negative feedback, whereas the feedback that is added is called positive feedback. The negative feedback is usually required for system stability, whereas positive feedback tends to make the system unstable. Unlike feedback control, open-loop control does not use the measured output to compute the control signal. The advantages of closed-loop control over open-loop control will be discussed in [Section 10.3](#).

Example 10.2: Closed-Loop Transfer Function

Reconsider the cart position control system in Example 10.1. The transfer functions of the plant (combining the cart and the DC motor), the controller, and the sensor are

$$G(s) = \frac{3.778}{s^2 + 16.883s}, \quad C(s) = 85, \quad H(s) = 1$$

Derive the closed-loop transfer functions $Y(s)/R(s)$ and $E(s)/R(s)$.

Solution

Using the result presented in [Section 4.5](#), the equivalent transfer function for a negative feedback control system is

$$\frac{Y(s)}{R(s)} = \frac{C(s)G(s)}{1 + C(s)G(s)H(s)}$$

Substituting the given transfer functions results in

$$\frac{Y(s)}{R(s)} = \frac{85 \left(\frac{3.778}{s^2 + 16.883s} \right)}{1 + 85 \left(\frac{3.778}{s^2 + 16.883s} \right)} = \frac{321.13}{s^2 + 16.883s + 321.13}$$

Since $E(s) = R(s) - Y(s)$, we have

$$\frac{E(s)}{R(s)} = \frac{R(s) - Y(s)}{R(s)} = 1 - \frac{Y(s)}{R(s)}$$

Inserting $Y(s)/R(s)$ gives

$$\frac{E(s)}{R(s)} = 1 - \frac{321.13}{s^2 + 16.883s + 321.13} = \frac{s^2 + 16.883s}{s^2 + 16.883s + 321.13}$$

Note that the system in Example 10.1 is designed to track a reference signal. This type of control system is called a tracking or *servo* system, in which the reference signal usually varies with time. If the reference signal is constant, usually zero, and the system is designed to hold an output steady against unknown disturbances, then the control system is known as a regulator.

PROBLEM SET 10.1

1. Draw a block diagram for the feedback control of a liquid-level system, which consists of a valve with a control knob (0%–100%) and a liquid-level sensor. Clearly label essential components and signals.
2. Draw a block diagram for the feedback control of a single-link robot arm system, which consists of a DC motor to produce the driving force and an encoder to measure the joint angle. Clearly label essential components and signals.
3. Determine the transfer functions $U(s)/E(s)$, $Y(s)/R(s)$, and $E(s)/R(s)$ in Figure 10.4.
4. The block diagram in Figure 10.5 represents a rocket attitude control system. Determine the transfer function $\Theta(s)/\Theta_r(s)$.

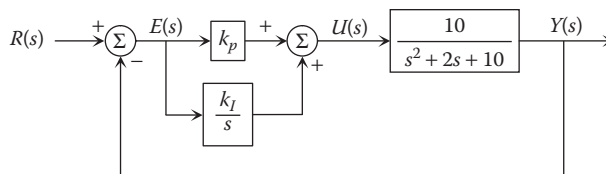


FIGURE 10.4

Problem 3.

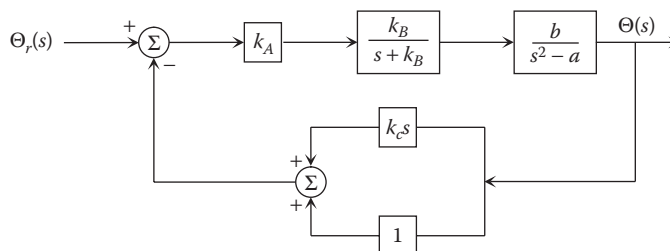


FIGURE 10.5

Problem 4.

5.  Consider the control system in Example 10.2. Build a Simulink block diagram to simulate reference tracking control, in which the signal $R(s)$ is a sine wave with a magnitude of 0.1 m and a frequency of 2 rad/s. Show the actual position response and the reference signal in the same scope.
6. Reconsider the control system in Example 10.2.
 - a. Convert the transfer function $G(s) = Y(s)/U(s)$ to a differential equation of $y(t)$.
 - b.  Using the differential equation obtained in Part (a) to represent the plant, build a Simulink block diagram to simulate regulation control, in which the reference signal $R(s)$ is zero. Assume that the initial conditions are $y(0) = 0.1$ m and $\dot{y}(0) = 0$ m/s.

10.2 Stability and Performance

Stability and performance are two important subjects in control. Generally, before designing a controller for a dynamic system, control designers check the stability and performance of the uncontrolled system. Then they come up with reasonable control design objectives from the perspective of the stability and performance. After a controller is designed, the stability and performance of the closed-loop control system are verified to meet the design objectives. In this section, we introduce the stability condition, performance specifications in time domain and frequency domain, and system identification.

10.2.1 Stability of Linear Time-Invariant Systems

Intuitively, a system is stable if its transient response decays and is unstable if it diverges. Thus, we can determine the stability of a system by solving and plotting the transient response of the system. However, this is not the only way to determine the system stability.

Consider a linear time-invariant system, whose transfer function is given by

$$G(s) = \frac{b(s)}{a(s)} = \frac{b_0 s^m + b_1 s^{m-1} + \cdots + b_{m-1} s + b_m}{s^n + a_1 s^{n-1} + \cdots + a_{n-1} s + a_n} \quad (10.3)$$

Assume that the numerator and denominator polynomials, $b(s)$ and $a(s)$, have no common factors. Setting the denominator of the transfer function equal to zero leads to the characteristic equation

$$a(s) = s^n + a_1 s^{n-1} + \cdots + a_{n-1} s + a_n = 0 \quad (10.4)$$

The roots of the characteristic equation are called the poles (see [Section 2.3](#)), which are complex and can be defined in terms of real and imaginary parts, $p = \sigma + j\omega$. A complex s -plane is usually sketched to show pole locations.

The following is the condition for stability: a linear time-invariant system is said to be stable if all the poles of its transfer function have negative real parts, and is unstable otherwise. In terms of the pole locations in the s -plane, a linear time-invariant system is stable if all the poles of the system are inside the left-half s -plane, that is, $\sigma < 0$. If even a single pole is in the right-half s -plane, the system is unstable. Thus, the imaginary $j\omega$ -axis, that is, $\sigma = 0$, is the stability boundary. If the system has nonrepeated poles on the $j\omega$ -axis, then the system is marginally stable. If the system has repeated poles on the $j\omega$ -axis, then the system is unstable. The roots of the numerator of the transfer function are called zeros, and are not related to the stability of the system.

Essentially, the transient response of a linear time-invariant system is associated with its pole locations in the s -plane. For example, consider a first-order system whose transfer function is $Y(s)/U(s) = 1/(s - p)$. Assume that the input is an impulse function. Inverse Laplace transformation yields the response $y(t) = e^{pt}$. The transient response e^{pt} approaches zero if and only if the real part of the pole p is negative. This simple example explains the reason why poles can be used to determine stability.

It is not an easy task to solve for the roots of a high-order characteristic equation by hand. Routh's stability criterion is a method for obtaining information about pole locations without solving for the poles. Consider the characteristic equation given in Equation 10.4. Routh's stability criterion consists of two conditions:

- A necessary (but not sufficient) condition for stability is that all the coefficients of the characteristic polynomial are positive.
- A necessary and sufficient condition for stability is that all the elements in the first column of the Routh array are positive.

If any of the coefficients in the characteristic polynomial is negative, we can conclude that the system is unstable by applying the first condition. If all the coefficients in the characteristic polynomial are positive, then we need to check the second condition by constructing a Routh array, as shown in Equation 10.5. For an n th order characteristic polynomial, the Routh array has $n + 1$ rows. The first two rows are obtained by arranging the coefficients of the characteristic polynomial, beginning with the first and second coefficients and followed by the even-numbered and odd-numbered coefficients, respectively. Starting from the third row, the elements are formed from the two previous rows using determinants, with two elements in the first column and the other two elements from successive columns. Equation 10.6 shows how to compute the elements in the third and fourth rows. The rest of the rows can be obtained in a manner similar to rows 3 and 4.

$$\begin{array}{rcll}
 s^n & : & 1 & a_2 & a_4 & a_6 & \cdots \\
 s^{n-1} & : & a_1 & a_3 & a_5 & a_7 & \cdots \\
 s^{n-2} & : & b_1 & b_2 & b_3 & \cdots & \\
 s^{n-3} & : & c_1 & c_2 & c_3 & \cdots & \\
 \vdots & : & \vdots & \vdots & \vdots & & \\
 s^2 & : & * & * & & & \\
 s^1 & : & * & & & & \\
 s^0 & : & * & & & &
 \end{array} \tag{10.5}$$

$$\begin{aligned}
 b_1 &= \frac{-\det \begin{bmatrix} 1 & a_2 \\ a_1 & a_3 \end{bmatrix}}{a_1} = \frac{a_1 a_2 - a_3}{a_1}, & c_1 &= \frac{-\det \begin{bmatrix} a_1 & a_3 \\ b_1 & b_2 \end{bmatrix}}{b_1} = \frac{b_1 a_3 - a_1 b_2}{b_1} \\
 b_2 &= \frac{-\det \begin{bmatrix} 1 & a_4 \\ a_1 & a_5 \end{bmatrix}}{a_1} = \frac{a_1 a_4 - a_5}{a_1}, & c_2 &= \frac{-\det \begin{bmatrix} a_1 & a_5 \\ b_1 & b_3 \end{bmatrix}}{b_1} = \frac{b_1 a_5 - a_1 b_3}{b_1} \\
 b_3 &= \frac{-\det \begin{bmatrix} 1 & a_6 \\ a_1 & a_7 \end{bmatrix}}{a_1} = \frac{a_1 a_6 - a_7}{a_1}, & c_3 &= \frac{-\det \begin{bmatrix} a_1 & a_7 \\ b_1 & b_4 \end{bmatrix}}{b_1} = \frac{b_1 a_7 - a_1 b_4}{b_1}
 \end{aligned} \tag{10.6}$$

Example 10.3: Stability

The transfer function of a dynamic system is given by

$$G(s) = \frac{s+4}{s^5 + 2s^4 + 3s^3 + 8s^2 + 4s + 5}$$

Determine the stability of the system

- Using Routh's stability criterion without solving for the poles.
- Using MATLAB to solve for the poles.

Solution

- The characteristic equation is $s^5 + 2s^4 + 3s^3 + 8s^2 + 4s + 5 = 0$. The Routh array can be formed as follows.

$$\begin{array}{rcl}
 s^5: & 1 & 3 \quad 4 \\
 s^4: & 2 & 8 \quad 5 \\
 s^3: & -1 = \frac{-\begin{vmatrix} 1 & 3 \\ 2 & 8 \end{vmatrix}}{2} & 1.5 = \frac{-\begin{vmatrix} 1 & 4 \\ 2 & 5 \end{vmatrix}}{2} \quad 0 \\
 s^2: & 11 = \frac{-\begin{vmatrix} 2 & 8 \\ -1 & 1.5 \end{vmatrix}}{-1} & 5 = \frac{-\begin{vmatrix} 2 & 5 \\ -1 & 0 \end{vmatrix}}{-1} \quad 0 \\
 s^1: & \frac{43}{22} = \frac{-\begin{vmatrix} -1 & 1.5 \\ 11 & 5 \end{vmatrix}}{11} & 0 \\
 s^0: & 5 = \frac{-\begin{vmatrix} 11 & 5 \\ \frac{43}{22} & 0 \end{vmatrix}}{\frac{43}{22}} &
 \end{array}$$

Because the elements in the first column of the Routh array are not all positive, we conclude that the system is unstable.

- b.  One of two MATLAB commands can be used to solve for the poles, `pole` or `roots`.

```
>> num = [1 4]; den = [1 2 3 8 4 5];
>> sys = tf(num,den);
>> pole(sys)
```

The command `pole` returns the poles of the system: -2.16 , $0.31 \pm 1.65j$, and $-0.23 \pm 0.88j$. There are two poles with positive real parts, and thus, the system is unstable. We can also use the command `roots` to solve for the roots of the characteristic equation.

```
>> roots(den)
```

Routh's stability criterion was especially useful before the availability of mathematical and scientific computing software, such as MATLAB. However, it is still useful for determining the ranges of system parameters for stability (see Example 10.6 in [Section 10.3](#)). It should be pointed out that the study of stability discussed here is limited to only linear time-invariant systems. The study of stability for nonlinear and time-varying systems is very complex and is beyond the scope of this text.

10.2.2 Time-Domain Performance Specifications

Performance specifications are certain requirements associated with the response of the system. In the time domain, the requirements are usually given for the step response. Consider a second-order system whose transfer function is given in the standard form (see [Section 8.4](#))

$$\frac{Y(s)}{U(s)} = \frac{\omega_n^2}{s^2 + 2\zeta\omega_n s + \omega_n^2} \quad (10.7)$$

[Figure 10.6](#) shows a unit-step response, in which the vertical axis is normalized, so that the steady-state value is equal to 1. Four quantities are defined to specify the performance of the system: rise time (t_r), overshoot (M_p), peak time (t_p), and settling time (t_s).

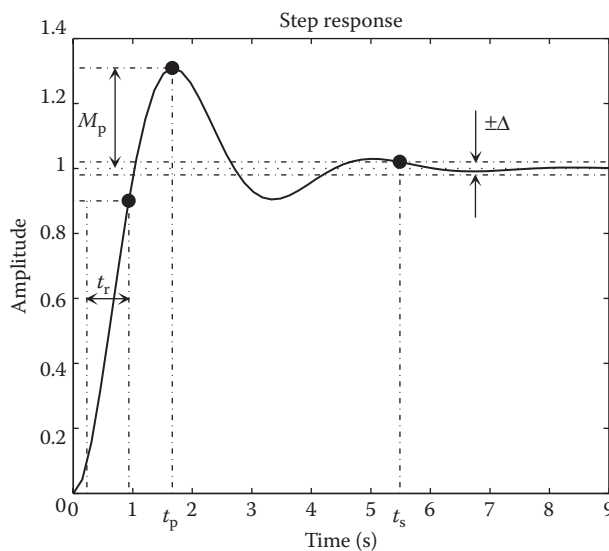


FIGURE 10.6

A unit-step response with time-domain performance specification indicated.

The rise time t_r is the time required for the response to rise from 10% to 90% of its steady-state value. The shorter the rise time, the faster the system reaches the vicinity of the steady-state value. The approximated relationship between t_r , the natural frequency ω_n , and the damping ratio ζ for a second-order system in standard form (see Equation 10.7) is given by

$$\omega_n t_r \approx 1.12 - 0.078\zeta + 2.230\zeta^2 \quad (10.8)$$

The overshoot M_p is the maximum amount of the system response exceeding the steady-state value divided by the steady-state value. It is usually expressed as a percentage value. The peak time t_p is the time it takes the response to reach the maximum value. As discussed in [Section 8.3](#), assuming zero initial conditions, the unit-step response for the second-order system in Equation 10.7 is

$$y(t) = 1 - e^{-\zeta\omega_n t} \left[\cos(\omega_d t) + \frac{\zeta}{\sqrt{1-\zeta^2}} \sin(\omega_d t) \right] \quad (10.9)$$

Differentiating $y(t)$ with respect to t and setting it equal to zero yields the peak time

$$t_p = \frac{\pi}{\omega_d} \quad (10.10)$$

Substituting Equation 10.10 into Equation 10.9 gives the value of y at the peak time. The overshoot M_p can be determined by computing $y(t_p) - 1$, as

$$M_p = e^{-\pi\zeta/\sqrt{1-\zeta^2}} \quad (10.11)$$

The detailed derivation will be left to the reader as an exercise. As shown in Equation 10.11, the overshoot of the step response is related to the damping of the system. [Figure 10.7](#) is the plot of M_p versus ζ . The larger the damping ratio is, the smaller the overshoot.

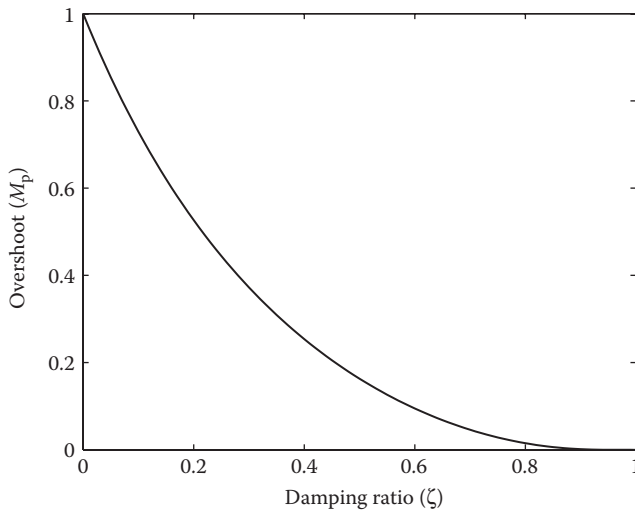


FIGURE 10.7
Plot of overshoot versus damping ratio for a second-order system.

The settling time t_s is the time required for the transient to decay to a small value such that $y(t)$ almost reaches the steady-state value. Note that Equation 10.9 can be rewritten as

$$y(t) = 1 - \frac{1}{\sqrt{1-\zeta^2}} e^{-\zeta\omega_n t} \sin(\omega_d t + \phi) \quad (10.12)$$

where

$$\phi = \tan^{-1} \frac{\sqrt{1-\zeta^2}}{\zeta} \quad (10.13)$$

It is observed from Equation 10.12 that the value of $y(t)$ is within the bounds

$$1 - \Delta = 1 - \frac{e^{-\zeta\omega_n t_s}}{\sqrt{1-\zeta^2}} \quad (10.14)$$

and

$$1 + \Delta = 1 + \frac{e^{-\zeta\omega_n t_s}}{\sqrt{1-\zeta^2}} \quad (10.15)$$

where Δ is a small value, such as 1%, 2%, and 5%. Thus, we have

$$t_s = -\frac{\ln(\Delta\sqrt{1-\zeta^2})}{\zeta\omega_n} \quad (10.16)$$

and for small damping,

$$t_s \approx -\frac{\ln \Delta}{\zeta\omega_n} \quad (10.17)$$

As indicated by Equations 10.8, 10.10, and 10.11, as well as 10.16 and 10.17, the time-domain performance specifications are related to the system parameters, specifically the undamped natural frequency ω_n and the damping ratio ζ . From Equation 10.7, the poles of a second-order system are also related to ω_n and ζ ,

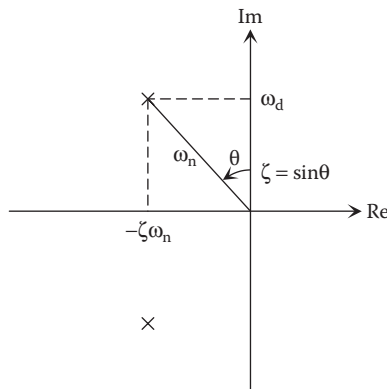
$$p_{1,2} = -\zeta\omega_n \pm j\omega_n\sqrt{1-\zeta^2} = -\zeta\omega_n \pm j\omega_d \quad (10.18)$$

which is a pair of complex conjugate poles for $0 \leq \zeta < 1$. [Figure 10.8](#) shows the plot of the poles in the s -plane. The two poles are equidistant from the origin with magnitude

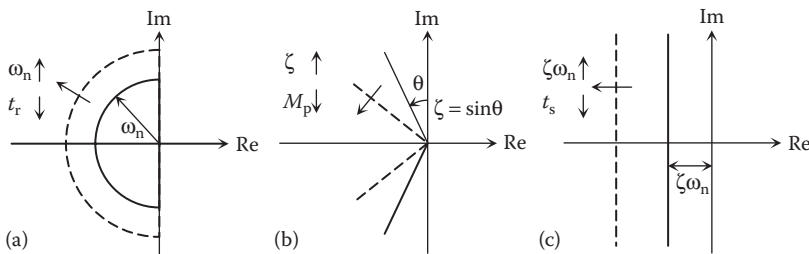
$$\sqrt{\text{Re}^2 + \text{Im}^2} = \sqrt{(-\zeta\omega_n)^2 + \omega_n^2(1-\zeta^2)} = \omega_n \quad (10.19)$$

and the angle between the pole and the imaginary axis satisfies

$$\sin \theta = \frac{|\text{Re}|}{\sqrt{\text{Re}^2 + \text{Im}^2}} = \frac{\zeta\omega_n}{\omega_n} = \zeta \quad (10.20)$$

**FIGURE 10.8**

Relationship between a pair of complex conjugate poles and system parameters.

**FIGURE 10.9**

Regions based on (a) rise time, (b) overshoot, and (c) settling time.

Therefore, the time-domain performance specifications are associated with the pole locations. In control design, one or more of these requirements are often specified to determine the allowable region for the poles in the s -plane. The three plots in Figure 10.9 show the regions based on the transient requirements t_r , M_p , and t_s , respectively.

Example 10.4: Time-Domain Performance Specifications

Consider a second-order system whose transfer function is in standard form as in Equation 10.7. Assume that the requirements for the system unit-step response are rise time $t_r \leq 0.1$ s, overshoot $M_p \leq 20\%$, and 2% settling time $t_s \leq 2$ s. Sketch the allowable region for the poles in the s -plane.

Solution

To express the damping ratio in terms of the overshoot, we take the natural logarithm of both sides of Equation 10.11,

$$\ln M_p = -\frac{\pi\zeta}{\sqrt{1-\zeta^2}}$$

Squaring both sides and solving for ζ ,

$$\zeta = \frac{|\ln M_p|}{\sqrt{\pi^2 + (\ln M_p)^2}}$$

As shown in both [Figures 10.7](#) and [10.9b](#), ζ is mono decreasing with respect to M_p . That is, ζ will increase if M_p decreases and vice versa. Thus, for $M_p \leq 20\%$, we have

$$\zeta \geq \frac{|\ln 0.2|}{\sqrt{\pi^2 + (\ln 0.2)^2}} = 0.46$$

The corresponding angle θ is

$$\theta = \sin^{-1} \zeta \geq \sin^{-1} 0.46 = 27^\circ$$

Equation 10.8 gives

$$\omega_n = \frac{1.12 - 0.078\zeta + 2.230\zeta^2}{t_r}$$

and ω_n increases when t_r decreases. Thus, for rise time $t_r \leq 0.1$ s and the smallest possible damping $\zeta = 0.46$, we have

$$\omega_n \geq \frac{1.12 - 0.078(0.46) + 2.230(0.46^2)}{0.1} = 15.56 \text{ rad/s}$$

Equation 10.16 indicates that

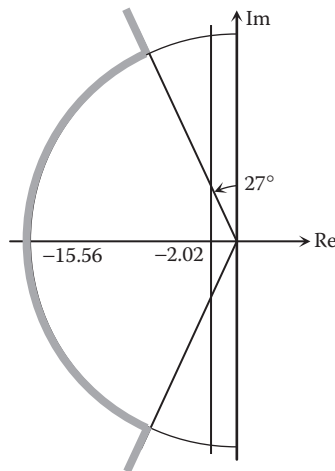
$$\zeta\omega_n = -\frac{\ln(\Delta\sqrt{1-\zeta^2})}{t_s}$$

and $\zeta\omega_n$ increases when t_s decreases. Thus, for 2% settling time $t_s \leq 2$ s and the smallest possible damping $\zeta = 0.46$, we have

$$\zeta\omega_n \geq -\frac{\ln(0.02\sqrt{1-0.46^2})}{2} = 2.02$$

The area to the left of the gray boundary shown in [Figure 10.10](#) is the allowable region for the poles in the s -plane, so that the three performance requirements are met. It can be seen that the requirement of settling time is automatically satisfied if the requirement of rise is met.

Note that Equations 10.8 through 10.17 are derived based on the assumption that the system has no zeros and has two complex poles. Thus, they do not provide precise design formulas for all systems. However, they can be used as qualitative guides to provide a starting point for design. The transient time response of the system is often checked to verify the time-domain specifications after control design.

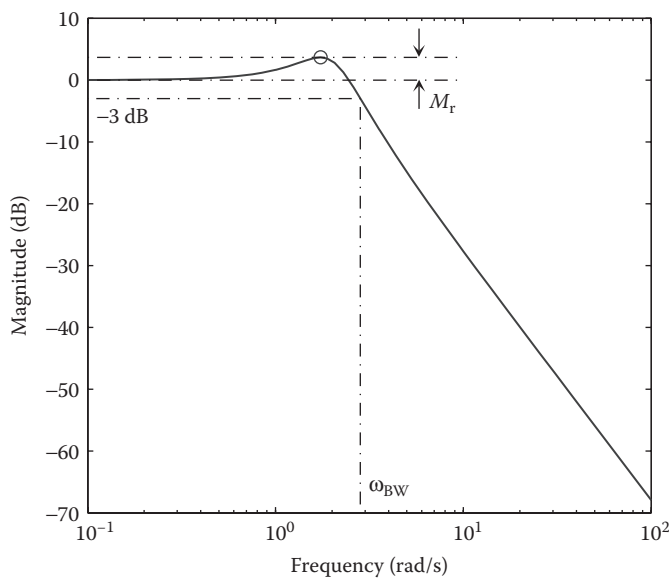
**FIGURE 10.10**

Allowable region for the poles in the s -plane.

10.2.3 Frequency-Domain Performance Specifications

System performance can also be specified in terms of frequency response. Figure 10.11 illustrates the ideal frequency response magnitude of a closed-loop control system. Two frequency-domain specifications are defined: bandwidth ω_{BW} and resonant peak M_r .

Assume that the transfer function of the closed-loop system is $Y(s)/R(s)$, where $Y(s)$ is the system output and $R(s)$ is the reference input. The bandwidth is defined as the frequency at which the magnitude of the closed-loop transfer function crosses -3 dB or 0.707 . Recall

**FIGURE 10.11**

An ideal frequency response with performance specifications indicated.

that the steady-state response of a linear system to sinusoidal excitations is called the system's frequency response. As shown in Figure 10.11, if the excitation frequency is lower than ω_{BW} , the magnitude $|Y(s)/R(s)|$ is close to 0 dB (or 1). This indicates that the system output follows the reference input. If the excitation frequency is higher than ω_{BW} , the magnitude $|Y(s)/R(s)|$ is reduced to a small value, and the system output no longer follows the reference input. The higher the bandwidth, the faster the reference signal the system can follow. Thus, the bandwidth is a measure of the speed of the response.

The resonant peak M_r was introduced previously in Section 8.4. It is defined as the maximum value of the magnitude of the frequency response. As shown in Section 8.4, the resonant peak is related to the damping of the system. The smaller the damping is, the higher the resonant peak. Compared with the time-domain performance specifications, the resonant peak M_r is similar to overshoot M_p , both of which are related to the damping ratio ζ , whereas the bandwidth ω_{BW} is similar to the rise time t_r , both related to the natural frequency ω_n .

10.2.4 Identification of Transfer Functions

As stated in Section 10.2.2, the time-domain performance specifications are related to system parameters, such as the undamped natural frequency ω_n and the damping ratio ζ . This relationship can be used to identify the transfer function of a dynamic system if it can be approximated as a second-order system.

Example 10.5: System Identification Using Time-Domain Performance Specifications

The unit-step response of a dynamic system is shown in Figure 10.12. Find the transfer function of the system if it can be approximated as $a/(s^2 + 2\zeta\omega_n s + \omega_n^2)$.

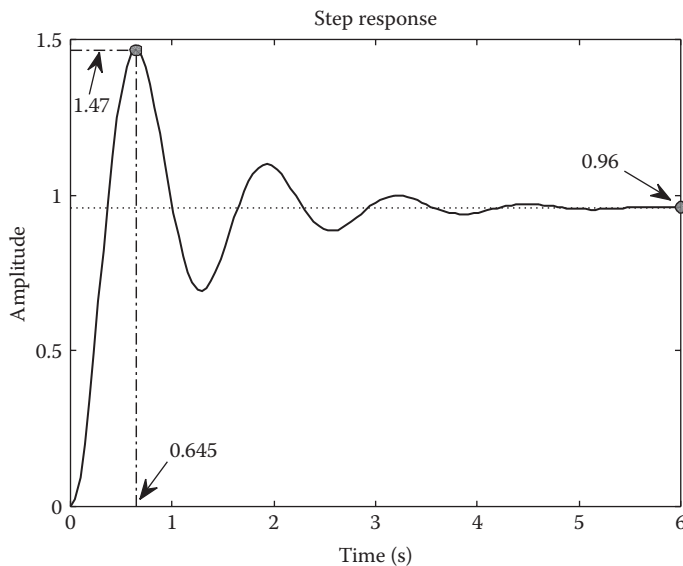


FIGURE 10.12

Unit-step response of a second-order system in Example 10.5.

Solution

As shown in [Figure 10.12](#), the unit-response of the system reaches a maximum value of 1.47 and the steady-state value is 0.96. Thus, the overshoot is

$$M_p = \frac{y_{\max} - y_{ss}}{y_{ss}} = \frac{1.47 - 0.96}{0.96} = 53.13\%$$

from which the damping ratio can be determined, and

$$\zeta = \frac{|\ln M_p|}{\sqrt{\pi^2 + (\ln M_p)^2}} = \frac{|\ln 0.5315|}{\sqrt{\pi^2 + (\ln 0.5315)^2}} = 0.20$$

Note that the peak time $t_p = 0.645$ s, which gives the damped natural frequency

$$\omega_d = \frac{\pi}{t_p} = \frac{\pi}{0.645} = 4.87 \text{ rad/s}$$

Thus, the undamped natural frequency is

$$\omega_n = \frac{\omega_d}{\sqrt{1 - \zeta^2}} = \frac{4.87}{\sqrt{1 - 0.20^2}} = 4.97 \text{ rad/s}$$

To determine the value of a , we apply the final-value theorem (see [Section 2.3](#)) to find the steady-state value of the unit-step response, that is,

$$y_{ss} = \lim_{s \rightarrow 0} s \frac{1}{s} \frac{a}{s^2 + 2\zeta\omega_n s + \omega_n^2} = \frac{a}{\omega_n^2}$$

which gives

$$a = y_{ss}\omega_n^2 = 0.96(4.97^2) = 23.71$$

Therefore, the transfer function of the system is $23.71/(s^2 + 1.98s + 24.70)$.

The frequency-domain performance specifications can also be used for system identification. More details will be given in [Section 10.6](#).

PROBLEM SET 10.2

1. The transfer function of a dynamic system is given by

$$G(s) = \frac{s + 1}{4s^4 + 5s^3 + 2s^2 + s + 6}$$

- a. Using Routh's stability criterion, determine the stability of the system.
- b.  Using MATLAB, solve for the poles of the system and verify the result obtained in Part (a).

2. The transfer function of a dynamic system is given by

$$G(s) = \frac{20s + 50}{s^3 + 10s^2 - 5s - 30}$$

- Using Routh's stability criterion, determine the stability of the open-loop system.
 - Suppose that a negative unity feedback is applied to this open-loop system. Using Routh's stability criterion, determine the stability of the resulting closed-loop system.
 - Using MATLAB, solve for the poles of the open and closed-loop systems and verify the results obtained in Parts (a) and (b).
3. The unit-step response for a second-order system $Y(s)/U(s) = \omega_n^2 / (s^2 + 2\zeta\omega_n s + \omega_n^2)$ is given by

$$y(t) = 1 - e^{-\zeta\omega_n t} \left[\cos(\omega_d t) + \frac{\zeta}{\sqrt{1-\zeta^2}} \sin(\omega_d t) \right]$$

Prove that the relationship between the overshoot M_p and the damping ratio is

$$M_p = e^{-\pi\zeta/\sqrt{1-\zeta^2}}$$

- Consider a second-order system $Y(s)/U(s) = \omega_n^2 / (s^2 + 2\zeta\omega_n s + \omega_n^2)$, which has two poles at $-4 \pm 4j$.
 - Determine the undamped natural frequency ω_n , damping ratio ζ , and damped natural frequency ω_d of the system.
 - Estimate the rise time t_r , overshoot M_p , peak time t_p , and 2% settling time t_s in the unit-step response for the system.
- Consider a second-order system $Y(s)/U(s) = \omega_n^2 / (s^2 + 2\zeta\omega_n s + \omega_n^2)$. Sketch the allowable region of the poles in the s -plane if the requirements for the system's unit-step response are $M_p \leq 10\%$ and rise time $t_r \leq 0.2$ s.
- Consider a second-order system $Y(s)/U(s) = \omega_n^2 / (s^2 + 2\zeta\omega_n s + \omega_n^2)$. Sketch the allowable region of the poles in the s -plane if the requirements for the system's unit-step response are $M_p \leq 25\%$, 5% settling time $t_s \leq 2$ s, and rise time $t_r \leq 0.5$ s.
- Consider a second-order system $Y(s)/U(s) = \omega_n^2 / (s^2 + 2\zeta\omega_n s + \omega_n^2)$.
 - Write a MATLAB m-file to plot the magnitude of the system's frequency response function for the following cases: $\omega_n = 2$ rad/s and $\zeta = 0.01, 0.1, 0.5$, and 1. Summarize the effects of the damping ratio on the frequency-domain performance.
 - Repeat Part (a) for the following cases: $\zeta = 0.1$ and $\omega_n = 1, 2$, and 6 rad/s. Summarize the effects of the natural frequency on the frequency-domain performance.
- The unit-step response of a dynamic system is shown in [Figure 10.13](#). The maximum value of the response is 1.1 and its steady-state value is 0.95. The response reaches 10% and 90% of its steady-state value at 0.048 and 0.21 s, respectively. Find the transfer function of the system if it can be approximated as $a/(s^2 + 2\zeta\omega_n s + \omega_n^2)$.

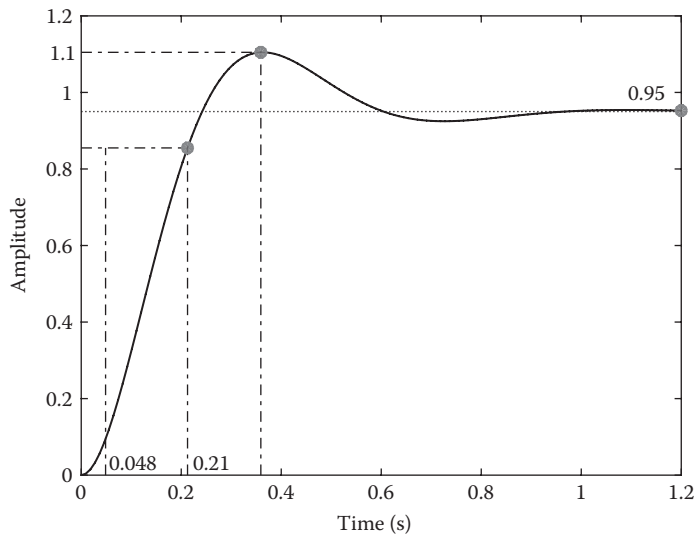


FIGURE 10.13
Problem 8.

10.3 Benefits of Feedback Control

As discussed in [Section 10.1](#), a closed-loop controller uses feedback to control the output of a plant. The input to a plant has an effect on its output, which is measured with a sensor and fed back to the controller, and then the computed control signal is used as the input to the plant, closing the loop. Feedback control has several advantages over open-loop control, such as stabilization, disturbance rejection, improved reference-tracking performance, and reduced sensitivity to parameter variations. In this section, we show these benefits one by one using MATLAB or Simulink. The controller in each discussion and example is assumed to be a gain denoted by K .

10.3.1 Stabilization

Consider a plant represented by a transfer function $G(s) = b(s)/a(s)$, where $a(s)$ and $b(s)$ are the denominator and numerator polynomials, respectively. Assume that $G(s)$ is unstable, which implies that not all the poles, or the roots of the characteristic equation $a(s) = 0$, have negative real parts. If an open-loop control system is implemented, as shown in [Figure 10.14](#), in which the output is $Y(s)$ and the input is $R(s)$, then the transfer function is

$$\frac{Y(s)}{R(s)} = KG(s) = \frac{Kb(s)}{a(s)} \quad (10.21)$$

It is obvious that the open-loop control system has the same poles as the plant $G(s)$, and thus, it is still unstable.

Now, we use a negative feedback control system. [Figure 10.15](#) shows the corresponding block diagram. The closed-loop transfer function is

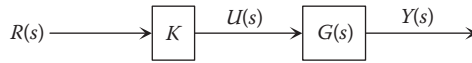


FIGURE 10.14
Open-loop control.

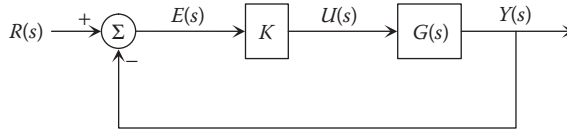


FIGURE 10.15
Closed-loop control.

$$\frac{Y(s)}{R(s)} = \frac{KG(s)}{1 + KG(s)} = \frac{Kb(s)}{a(s) + Kb(s)} \quad (10.22)$$

The characteristic equation of the closed-loop system is therefore $a(s) + Kb(s) = 0$. Properly choosing the control gain K can possibly make the closed-loop system stable. Routh's stability criterion introduced in [Section 10.2](#) is one way to determine the range of K such that the closed-loop control system is stable.

Example 10.6: Stabilization Using Feedback

Consider an unstable plant

$$G(s) = \frac{s+2}{s^3 + 4s^2 - 5s}$$

with feedback control, as shown in [Figure 10.15](#).

- Using Routh's stability criterion, determine the range of the control gain K for which the closed-loop system is stable.
- Use MATLAB commands to find the unit-step responses for open and closed-loop control. Assuming that the control gain is $K = 20$, compare the open- and closed-loop responses.

Solution

- Solving the characteristic equation of $G(s)$ yields the poles 0, 1, and -5 . The positive real pole, 1, indicates that the plant $G(s)$ is unstable. The same conclusion can also be made using the first condition of Routh's stability criterion, because the characteristic equation of $G(s)$ has one negative coefficient, -5 .

The closed-loop transfer function is

$$\frac{Y(s)}{R(s)} = \frac{K \frac{s+2}{s^3 + 4s^2 - 5s}}{1 + K \frac{s+2}{s^3 + 4s^2 - 5s}} = \frac{Ks + 2K}{s^3 + 4s^2 + (K-5)s + 2K}$$

The closed-loop characteristic equation is

$$s^3 + 4s^2 + (K-5)s + 2K = 0$$

for which we can construct the Routh array as follows:

$$\begin{array}{rcl} s^3: & 1 & K-5 \\ s^2: & 4 & 2K \\ s^1: & \frac{K}{2}-5 = \frac{4(K-5)-2K}{4} & 0 \\ s^0: & 2K & \end{array}$$

To make the closed-loop system stable, all elements in the first column of the Routh array must be positive. Therefore,

$$\begin{cases} \frac{K}{2}-5 > 0 \\ 2K > 0 \end{cases}$$

which leads to $K > 10$.

- b.  For the open-loop control system shown in Figure 10.14, the controller block is connected with the plant in series. The following is the MATLAB session.

```
>> G = tf([1 2],[1 4 -5 0]);
>> K = 20;
>> olp = K*G;
>> step(olp);
```

Figure 10.16 shows the unit-step response of the open-loop control system, which is obviously unstable because the response diverges. Therefore, the unstable plant cannot be stabilized using open-loop control.

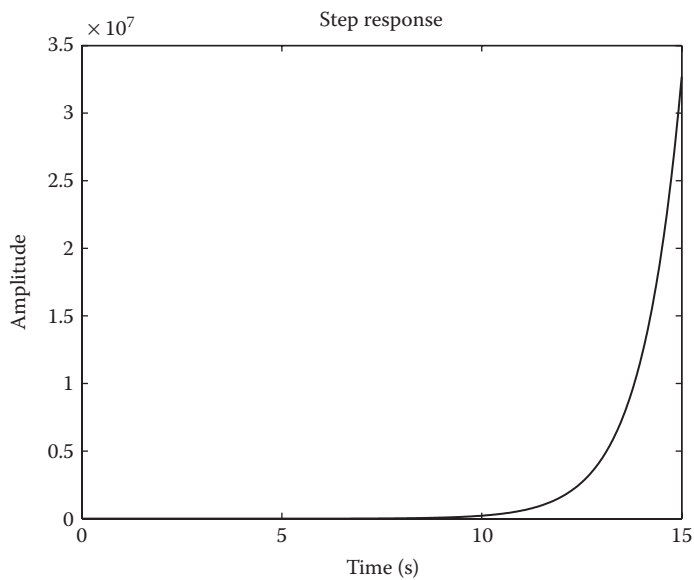
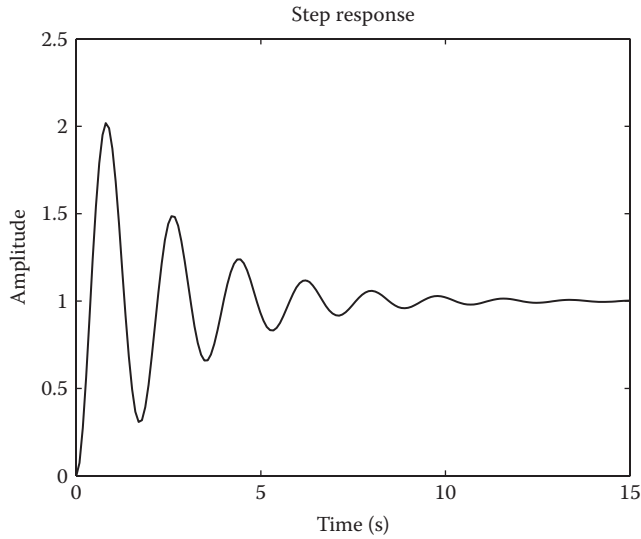


FIGURE 10.16

Unit-step response of the open-loop control system in Example 10.6.

**FIGURE 10.17**

Unit-step response of the closed-loop control system in Example 10.6.

As introduced in [Section 4.5](#), the MATLAB command `feedback` can be used to find the transfer function of the closed-loop system for the feedback control system in [Figure 10.15](#).

```
>> clp = feedback(K*G,1);
>> step(clp);
```

As observed in [Figure 10.17](#), the unit-step response of the closed-loop control system converges and the unstable plant is stabilized.

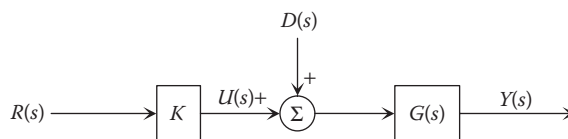
10.3.2 Disturbance Rejection

To compare the capabilities of the open-loop and the closed-loop control for disturbance rejection, assume that a disturbance is an input to the plant $G(s)$. [Figure 10.18](#) is the block diagram for the open-loop control, where

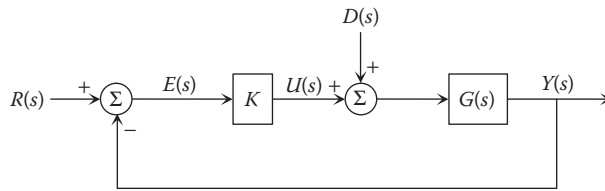
$$Y(s) = G(s)U(s) + G(s)D(s) = KG(s)R(s) + G(s)D(s) \quad (10.23)$$

Equation 10.23 shows that the output $Y(s)$ depends on the reference input $R(s)$ and the disturbance input $D(s)$. Letting $R(s) = 0$ yields the transfer function relating the disturbance $D(s)$ and the output $Y(s)$,

$$\frac{Y(s)}{D(s)} = G(s) \quad (10.24)$$

**FIGURE 10.18**

Open-loop control with disturbance input.

**FIGURE 10.19**

Closed-loop control with disturbance input.

Note that K does not appear in Equation 10.24 and has no control over the disturbance in the open-loop case.

Using closed-loop control as shown in Figure 10.19, we have

$$Y(s) = G(s)U(s) + G(s)D(s) = KG(s)[R(s) - Y(s)] + G(s)D(s) \quad (10.25)$$

Solving for $Y(s)$ gives

$$Y(s) = \frac{KG(s)}{1 + KG(s)} R(s) + \frac{G(s)}{1 + KG(s)} D(s) \quad (10.26)$$

The transfer function relating the disturbance $D(s)$ and the output $Y(s)$ is

$$\frac{Y(s)}{D(s)} = \frac{G(s)}{1 + KG(s)} \quad (10.27)$$

which indicates that K in the closed-loop case has control over the disturbance.

Example 10.7: Disturbance Rejection Using Feedback

Consider a plant whose transfer function is

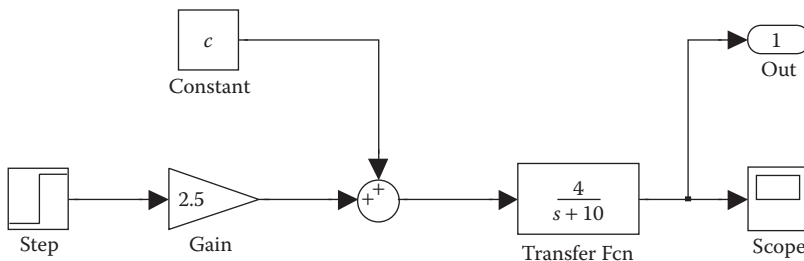
$$G(s) = \frac{4}{s + 10}$$

- Build a Simulink block diagram associated with Figure 10.18 to simulate open-loop control with disturbance input. Assume that the controller is $K = 2.5$, the disturbance is a constant of -1 , and the reference is a unit-step function with the step time at $t = 0$ s. Compare the steady-state values of the responses without and with the disturbance.
- Build a Simulink block diagram associated with Figure 10.19 to simulate closed-loop control with disturbance input. Assume that the controller is $K = 50$. The disturbance and the reference are the same as those in Part (a). Compare the steady-state values of the responses without and with the disturbance.

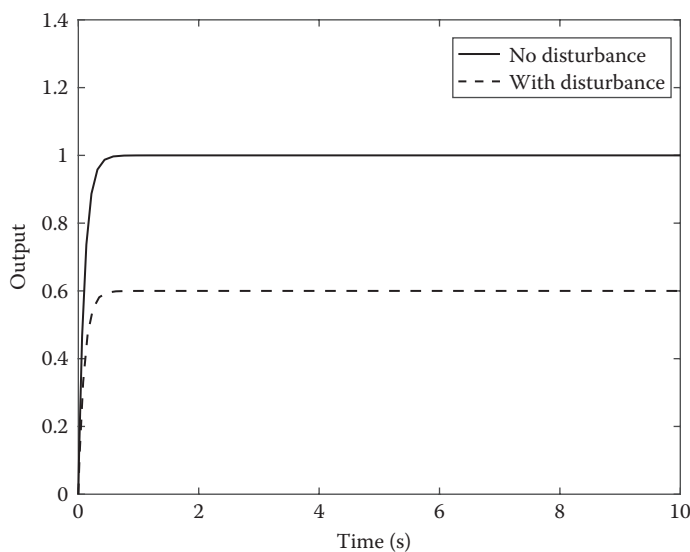
Solution

- a. The Simulink block diagram of the open-loop control is shown in Figure 10.20, in which the Constant block is used to represent the disturbance signal. We first set the Constant block to 0 and run the simulation. Then, we change the Constant block to -1 and rerun the simulation. The responses for those two cases are plotted in Figure 10.21. The steady-state value of the unit-step response is 1 without disturbance and 0.6 with disturbance.
- b. The Simulink block diagram of the closed-loop control is shown in Figure 10.22. The responses are plotted in Figure 10.23. The steady-state value of the unit-step response is 0.95 without disturbance and 0.93 with disturbance.

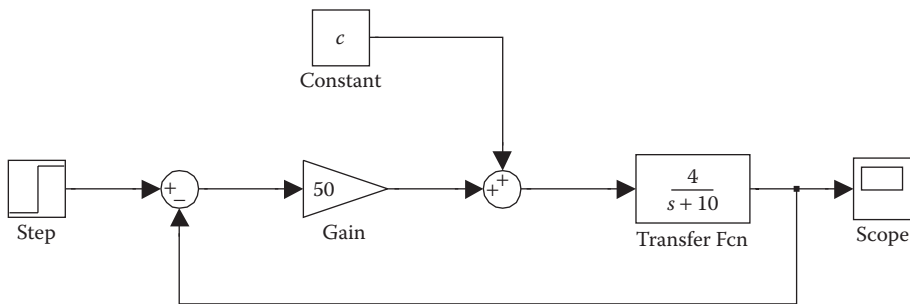
Although the closed-loop control is not as good as the open-loop control in the absence of disturbance, the error resulting from a constant disturbance can be made smaller in a closed-loop feedback system when compared with an open-loop system.

**FIGURE 10.20**

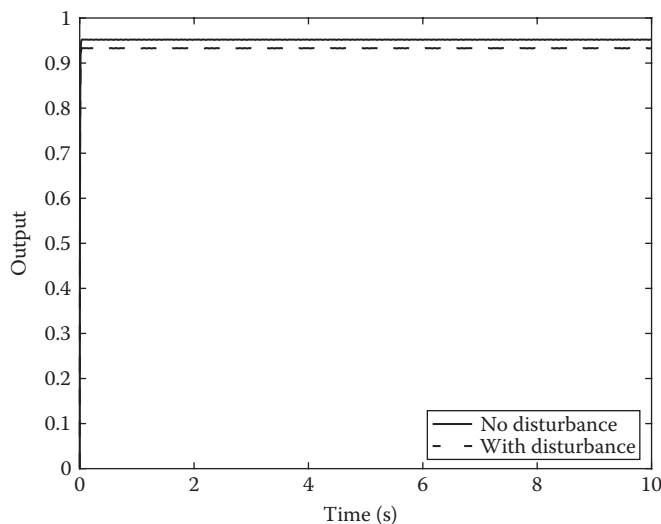
Simulink block diagram of open-loop control with disturbance input.

**FIGURE 10.21**

Unit-step responses of open-loop control without- and with disturbance input.

**FIGURE 10.22**

Simulink block diagram of closed-loop control with disturbance input.

**FIGURE 10.23**

Unit-step responses of closed-loop control without- and with disturbance input.

10.3.3 Reference Tracking

As introduced in [Section 10.1](#), there are two types of control systems: regulators and servos. The former ones are designed for disturbance rejection and the latter are designed for reference tracking. With feedback, a control system can achieve improved reference tracking performance. To show this, let us consider the closed-loop control case in [Figure 10.19](#), in which the disturbance $D(s)$ is now set to zero. As a result, the closed-loop transfer function of the system is

$$\frac{Y(s)}{R(s)} = \frac{KG(s)}{1 + KG(s)} \quad (10.28)$$

A typical frequency response plot of the closed-loop system was given in [Figure 10.11](#). As we discussed in [Section 10.2](#), the output follows the reference input when $|Y(s)/R(s)| \approx 1$. Equation 10.28 implies that $Y(s)$ is approximately equal to $R(s)$ if the magnitude $|KG(s)| \gg 1$. In general, this can be done by increasing the value of the control gain K . Thus, a large control gain can reduce the steady-state error of the response.

Example 10.8: Reference Tracking Using Feedback

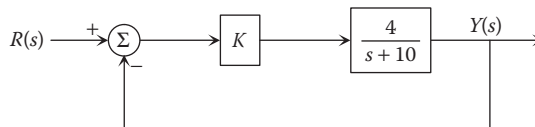
Consider the closed-loop control system shown in Figure 10.24.

- Using MATLAB, plot the unit-step responses of the system for the following values of K : 5, 50, and 500.
- Compute the steady-state errors for the different values of K .

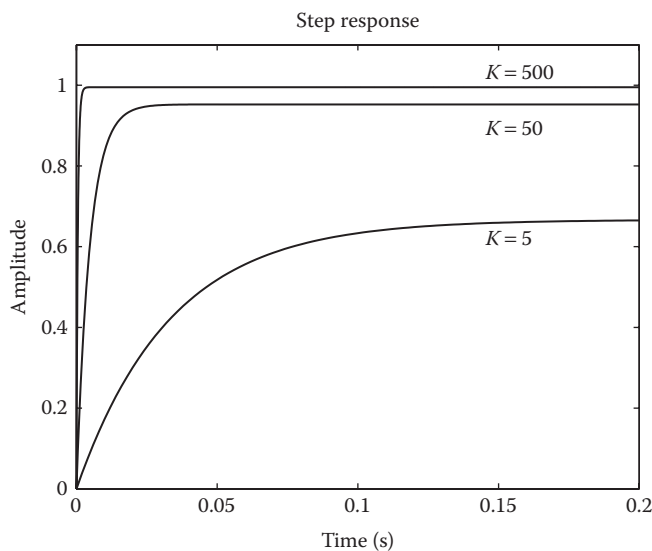
Solution

- The following is the MATLAB script. The unit-step responses of the system for the three different values of the control gain are shown in Figure 10.25. It is observed that the steady-state error becomes smaller as the control gain increases.

```
K = [5 50 500];
G = tf([4],[1 10]);
for i = 1:length(K)
    clp = feedback(K(i)*G,1);
    step(clp);
    hold on;
end
```

**FIGURE 10.24**

A reference tracking control system with feedback.

**FIGURE 10.25**

Unit-step responses for the system in Figure 10.24, with different values of K .

- b. The value of the steady-state error can be computed by applying the final-value theorem. The closed-loop transfer function $E(s)/R(s)$ is

$$\frac{E(s)}{R(s)} = \frac{1}{1 + KG(s)}$$

For a unit-step input, we have

$$e_{ss} = \lim_{s \rightarrow 0} sE(s) = \lim_{s \rightarrow 0} sR(s) \frac{E(s)}{R(s)} = \lim_{s \rightarrow 0} s \frac{1}{s} \frac{1}{1 + KG(s)} = \lim_{s \rightarrow 0} \frac{s + 10}{s + 10 + 4K} = \frac{10}{10 + 4K}$$

The steady-state errors for $K = 5, 50$, and 500 are $0.3333, 0.0476$, and 0.005 , respectively. It should be pointed out that a large control gain may also result in unsatisfactory transient response and even destabilize the system. Details will be discussed in [Section 10.4](#).

10.3.4 Sensitivity to Parameter Variations

For model-based control, a controller implemented for a practical system is designed based on a mathematical model of the system. Therefore, it is important to obtain a precise model of the plant for control design. However, modeling errors do exist due to uncertainties in system parameters. For example, the dynamic behavior of a mass–spring system depends on the values of the mass and stiffness. The values used for modeling might be different from the actual values due to inevitable measurement errors from the start or slight parameter changes caused by external effects. To maintain control performance, a controller should be insensitive to parameter changes.

Consider a function f that depends on a parameter a . Denote the change in the parameter as δa . If the change in f due to the parameter change is δf , then the sensitivity of the function f with respect to the parameter a is defined as

$$S_a^f = \frac{\delta f / f}{\delta a / a} \quad (10.29)$$

where the first-order variation δf is proportional to the derivative df/da and is given by

$$\delta f = \frac{df}{da} \delta a \quad (10.30)$$

Thus, the sensitivity S_a^f can be written as

$$S_a^f = \frac{a}{f} \frac{df}{da} \quad (10.31)$$

To compare the sensitivity of open-loop control with that of closed-loop control, we assume that one or more parameters in the plant change. Without loss of generality, the disturbance is assumed to be zero. For the open-loop control shown in [Figure 10.18](#), denote the transfer function from the reference signal to the system output as $T_o(s)$, which is $KG(s)$. Applying Equation 10.31 gives the sensitivity of the open-loop control system $T_o(s)$ to the plant $G(s)$

$$S_G^{T_o} = \frac{G}{T_o} \frac{dT_o}{dG} = \frac{G}{KG} \frac{d(KG)}{dG} = \frac{G}{KG} \frac{K dG}{dG} = 1 \quad (10.32)$$

which implies that open-loop control is very sensitive to the parameter variations in the plant. For example, a 5% error in the plant would yield the same percentage error in the open-loop transfer function.

For the closed-loop control case as shown in Figure 10.19, in which the disturbance is still set as zero, denote the closed-loop transfer function as $T_c(s)$. Applying Equation 10.31 gives

$$S_G^{T_c} = \frac{G}{T_c} \frac{dT_c}{dG} \quad (10.33)$$

The closed-loop transfer function $T_c(s)$ is given by Equation 10.28. Thus, we have

$$\frac{dT_c}{dG} = \frac{d}{dG} \left(\frac{KG}{1+KG} \right) = \frac{K(1+KG) - KG(K)}{(1+KG)^2} = \frac{K}{(1+KG)^2} \quad (10.34)$$

Substituting it into Equation 10.33 yields

$$S_G^{T_c} = \frac{G(1+KG)}{KG} \frac{K}{(1+KG)^2} = \frac{1}{1+KG} \quad (10.35)$$

which can be made considerably smaller than 1 by adjusting the controller gain K . Thus, the overall transfer function in feedback control is less sensitive to variations in the plant gain compared with the one in open-loop control.

Example 10.9: Sensitivity to Parameter Variations

Consider a mass–damper–spring system $G(s) = Y(s)/U(s) = 1/(ms^2 + bs + k)$, where $m = 1$ kg, $b = 8$ N·s/m, and $k = 40$ N/m.

- Assume that the system is controlled in an open-loop control system with a controller $K = 40$. Determine the steady-state value of the response to a unit-step input. If the spring stiffness is actually 50 N/m, recalculate the steady-state value of the response and determine the fractional change in the steady-state value.
- Repeat Part (a) assuming that the system is controlled in a feedback control system with a controller $K = 2000$.

Solution

- In the open-loop case, the steady-state value of the response to a unit-step input is

$$y_{ss} = \lim_{s \rightarrow 0} sR(s) \frac{Y(s)}{R(s)} = \lim_{s \rightarrow 0} s \frac{1}{s} KG(s) = \lim_{s \rightarrow 0} \frac{40}{s^2 + 8s + 40} = 1$$

If the spring stiffness is 50 N/m, then the steady-state value becomes

$$y_{ss} = \lim_{s \rightarrow 0} sR(s) \frac{Y(s)}{R(s)} = \lim_{s \rightarrow 0} s \frac{1}{s} KG(s) = \lim_{s \rightarrow 0} \frac{40}{s^2 + 8s + 50} = 0.8$$

Note that y_{ss} is reduced by 20%, which is the same as the relative error in k .

- In the closed-loop case, the steady-state value of the response to a unit-step input is

$$y_{ss} = \lim_{s \rightarrow 0} sR(s) \frac{Y(s)}{R(s)} = \lim_{s \rightarrow 0} s \frac{1}{s} \frac{KG(s)}{1+KG(s)} = \lim_{s \rightarrow 0} \frac{2000}{s^2 + 8s + 2040} = 0.9804$$

If the spring stiffness is 50 N/m, then the steady-state value becomes

$$y_{ss} = \lim_{s \rightarrow 0} sR(s) \frac{Y(s)}{R(s)} = \lim_{s \rightarrow 0} s \frac{1}{s} \frac{KG(s)}{1 + KG(s)} = \lim_{s \rightarrow 0} \frac{2000}{s^2 + 8s + 2050} = 0.9756$$

Note that y_{ss} is reduced by 0.49% only, which is much less than the relative error in k . Although the closed-loop control is not as good as the open-loop control in the absence of modeling error, the tracking error resulting from parameter variations can be made smaller in a closed-loop feedback system compared with an open-loop system.

PROBLEM SET 10.3

- Consider the feedback system shown in Figure 10.26.
 - Using Routh's stability criterion, determine the range of the control gain K for which the closed-loop system is stable.
 - Use MATLAB commands to find the unit-step responses for open and closed-loop control. Assume that the control gain is $K = 30$. Compare the open- and closed-loop responses.
- Consider the feedback system shown in Figure 10.27. Using Routh's stability criterion, determine the range of the control gain K for which the closed-loop system is stable.
- Reconsider Example 10.7. Using the final-value theorem, verify the steady-state errors to a unit-step input for open and closed-loop control without- and with disturbance.
- A stable system can be classified as a system type, which is defined as the degree of the polynomial for which the steady-state error is a non-zero finite constant. For instance, if the error to a step input, which is a polynomial of zero degree, is a non-zero finite constant, then such a system is called type 0, and so on. Consider the system in Figure 10.28.
 - Compute the following steady-state errors for: (1) a unit-step reference input, (2) a unit-ramp reference input $u = t$, and (3) a parabolic reference input $u = 0.5t^2$.
 - Determine the system type.

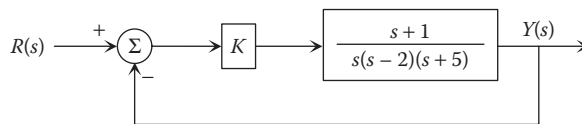


FIGURE 10.26
Problem 1.

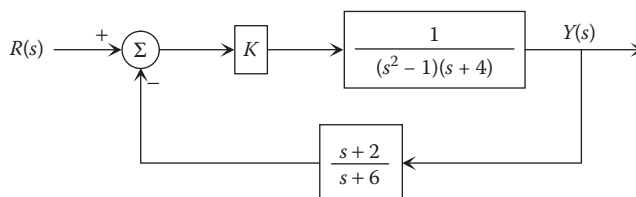


FIGURE 10.27
Problem 2.

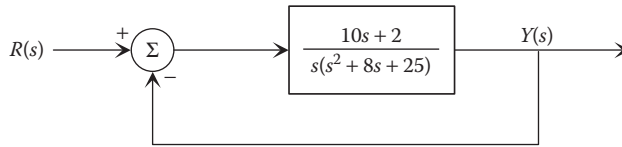


FIGURE 10.28
Problem 4.

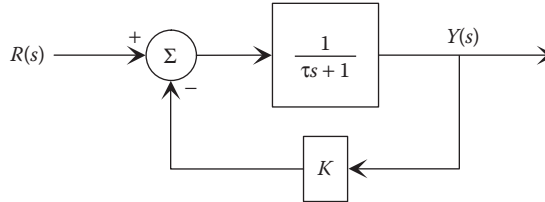


FIGURE 10.29
Problem 6.

5. Reconsider Example 10.9. Build Simulink block diagrams to simulate open and closed-loop control with parameter variations. Verify the steady-state response values y_{ss} obtained in Example 10.9.
6. Consider the feedback control system shown in [Figure 10.29](#).
 - a. Compute the sensitivity of the closed-loop transfer function to changes in the parameter τ .
 - b. Compute the sensitivity of the closed-loop transfer function to changes in the parameter K .
 - c. Assuming $\tau = 1$ and $K = 1$, use MATLAB to plot the magnitude of each of the sensitivity functions for $s = j\omega$. Use the logarithmic scale for the y -axis. Comment on the effect of parameter variations in τ and K for different driving frequencies ω .

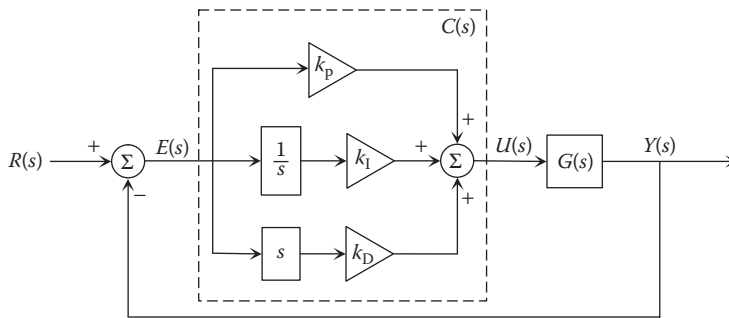
10.4 Proportional–Integral–Derivative Control

A proportional–integral–derivative (PID) controller is a generic feedback control structure widely used in industries. The PID controller involves three terms. The proportional term determines the control signal based on the current error, the integral term determines the control signal based on the integral of error, and the derivative term determines the control signal based on the derivative of error. The expression of the PID controller in the time domain is given by

$$u(t) = k_p e(t) + k_i \int_0^t e(\tau) d\tau + k_D \frac{de(t)}{dt} \quad (10.36)$$

where:

- k_p is the proportional control gain
- k_i is the integral control gain
- k_D is the derivative control gain

**FIGURE 10.30**

A block diagram of PID feedback control.

Taking the Laplace transform of Equation 10.36 yields the transfer function of the PID controller as

$$\frac{U(s)}{E(s)} = k_p + \frac{k_I}{s} + k_D s \quad (10.37)$$

Figure 10.30 shows a block diagram of feedback control using a PID controller denoted by $C(s)$. In this section, we discuss the advantages and disadvantages of PID control.

10.4.1 Proportional Control

For proportional feedback control, the controller transfer function is

$$\frac{U(s)}{E(s)} = k_p \quad (10.38)$$

Note that the controller structures in the previous sections represent the most basic proportional feedback control. As discussed in Section 10.3, a high proportional control gain can result in smaller steady-state error. However, if k_p is made too large, the closed-loop system may have reduced damping and even become unstable.

Example 10.10: Proportional Control

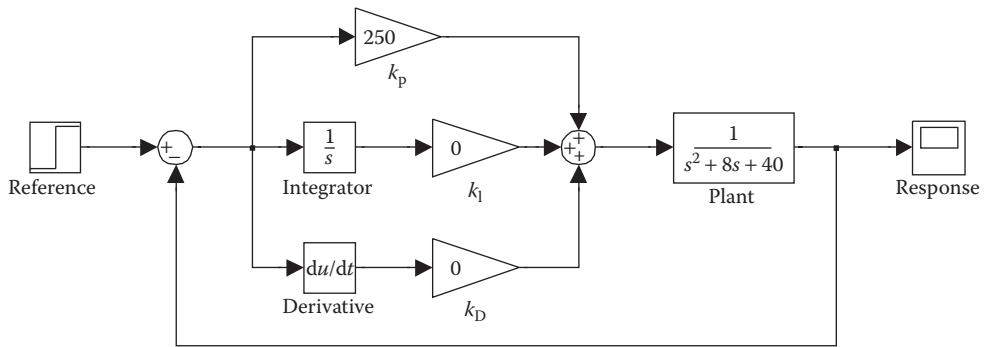
Consider the mass–damper–spring system in Example 10.9,

$$G(s) = \frac{1}{s^2 + 8s + 40}$$

Use Simulink to build a block diagram for proportional feedback control. Find the unit-step responses for $k_p = 25$, 250, and 2000. Discuss the effects of the proportional feedback on the unit-step response.

Solution

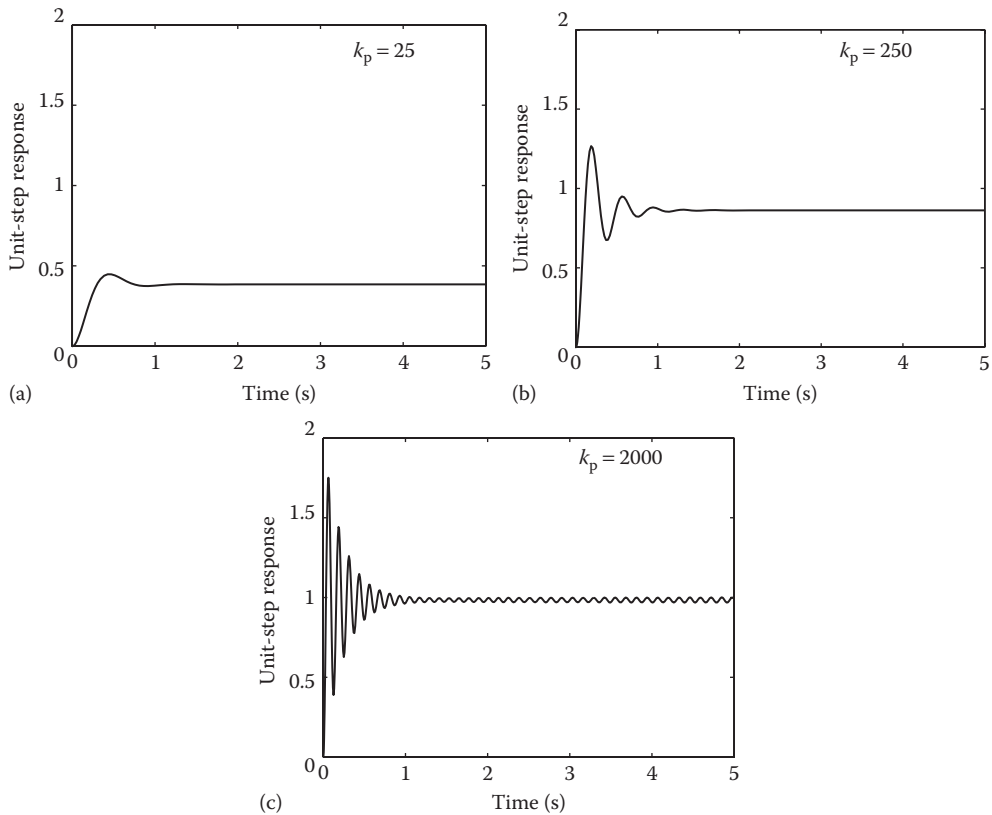
The Simulink block diagram is shown in Figure 10.31. Note that Figure 10.31 also includes integral and derivative control loops, which will be used later in this section. Set k_I and k_D to zero for the current example. Run the simulation for $k_p = 25$, 250, and 2000.

**FIGURE 10.31**

Simulink block diagram for PID feedback control.

Figure 10.32 illustrates the effects of proportional feedback control. It is readily observed that as k_p increases, the steady-state error to the unit-step input becomes smaller. Applying the final-value theorem gives

$$e_{ss} = \lim_{s \rightarrow 0} sR(s) \frac{E(s)}{R(s)} = \lim_{s \rightarrow 0} s \frac{1}{s} \frac{1}{1 + k_p G(s)} = \frac{40}{40 + k_p}$$

**FIGURE 10.32**

Responses of proportional control for (a) $k_p = 25$, (b) $k_p = 250$, and (c) $k_p = 2000$.

Substituting the three different proportional control gains $k_p = 25, 250$, and 2000 yields steady-state errors of $e_{ss} = 0.62, 0.14$, and 0.02 . For this particular system, the steady-state error will approach zero as k_p increases. However, it will never be zero.

On the other hand, the larger proportional gain results in less satisfactory oscillatory response. This is caused by reduced damping. Note that the characteristic equation of the closed-loop system with proportional control is

$$s^2 + 8s + 40 + k_p = 0$$

For a second-order system, the coefficients in the characteristic equation are related to the natural frequency and the damping ratio of the system, that is,

$$8 = 2\zeta\omega_n, \quad 40 + k_p = \omega_n^2$$

Obviously, if k_p is made large, the natural frequency is increased. However, the damping ratio becomes smaller. This leads to a faster response, but with a bigger overshoot and much more severe oscillations.

10.4.2 Proportional–Integral Control

As seen in Example 10.10, increasing the proportional gain k_p can reduce steady-state error, but cannot achieve zero steady-state error. Adding an integral term to the controller in Equation 10.38 results in a proportional–integral (PI) controller

$$\frac{U(s)}{E(s)} = k_p + \frac{k_I}{s} \quad (10.39)$$

If PI control is used in the previous example, the steady-state error becomes

$$e_{ss} = \lim_{s \rightarrow 0} s \frac{1}{s} \frac{1}{1 + \left(k_p + \frac{k_I}{s}\right) G(s)} = \lim_{s \rightarrow 0} \frac{s}{s + (k_p s + k_I) G(s)} \quad (10.40)$$

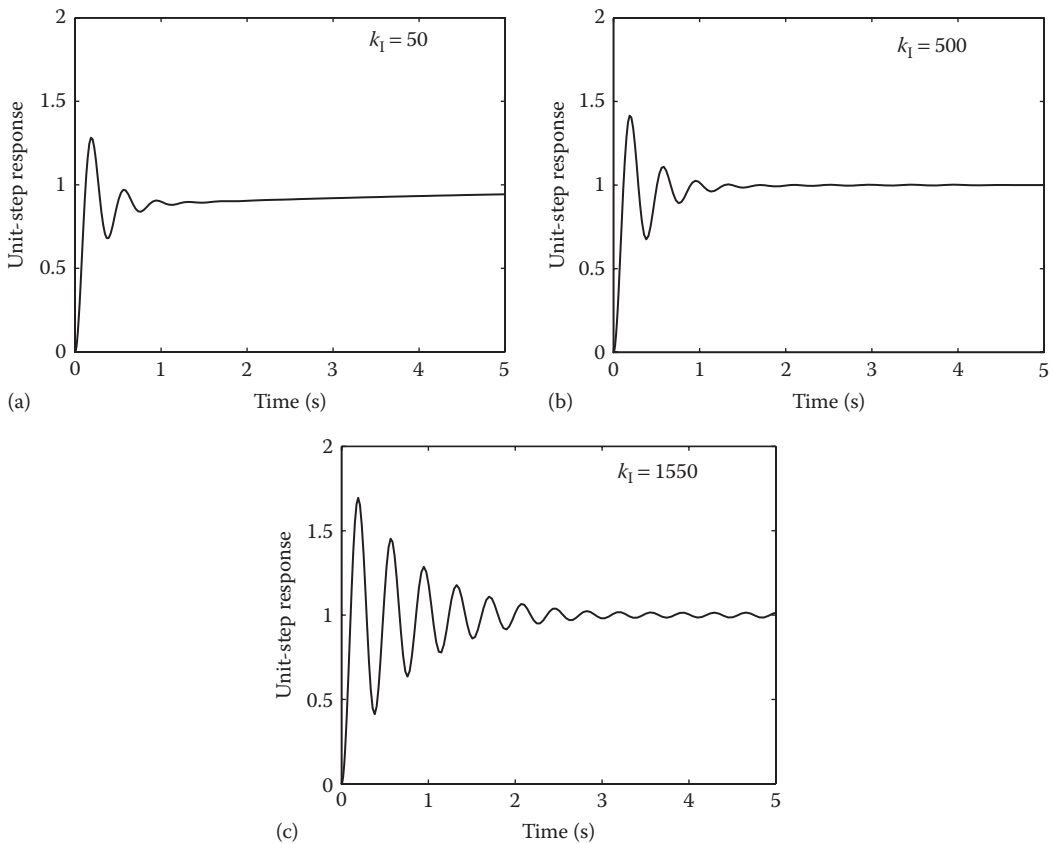
Substituting the transfer function $G(s)$ given in Example 10.10, we can obtain the result by evaluating the limit in Equation 10.40. The system with PI control has a zero steady-state error. Thus, the primary reason to introduce the integral control is to reduce, or possibly eliminate, the steady-state error.

Example 10.11: Proportional–Integral Control

Use the Simulink block diagram built in Example 10.10 to find the unit-step responses for $k_I = 50, 500$, and 1550 . Set $k_p = 250$ and $k_D = 0$. Discuss the effects of the integral term on the unit-step response.

Solution

Figure 10.33 shows the unit-step responses of PI feedback control for $k_I = 50, 500$, and 1550 . The system response will achieve zero for each case if the simulation time is long enough. However, a large integral control gain results in lightly damped oscillations.

**FIGURE 10.33**

Responses of PI control for (a) $k_I = 50$, (b) $k_I = 500$, and (c) $k_I = 1550$.

10.4.3 Proportional–Integral–Derivative Control

The final term in a PID controller represents the derivative control. The complete three-term controller is given by Equation 10.37. The main reason to introduce the derivative control is to increase the damping and thus to improve the stability of the system.

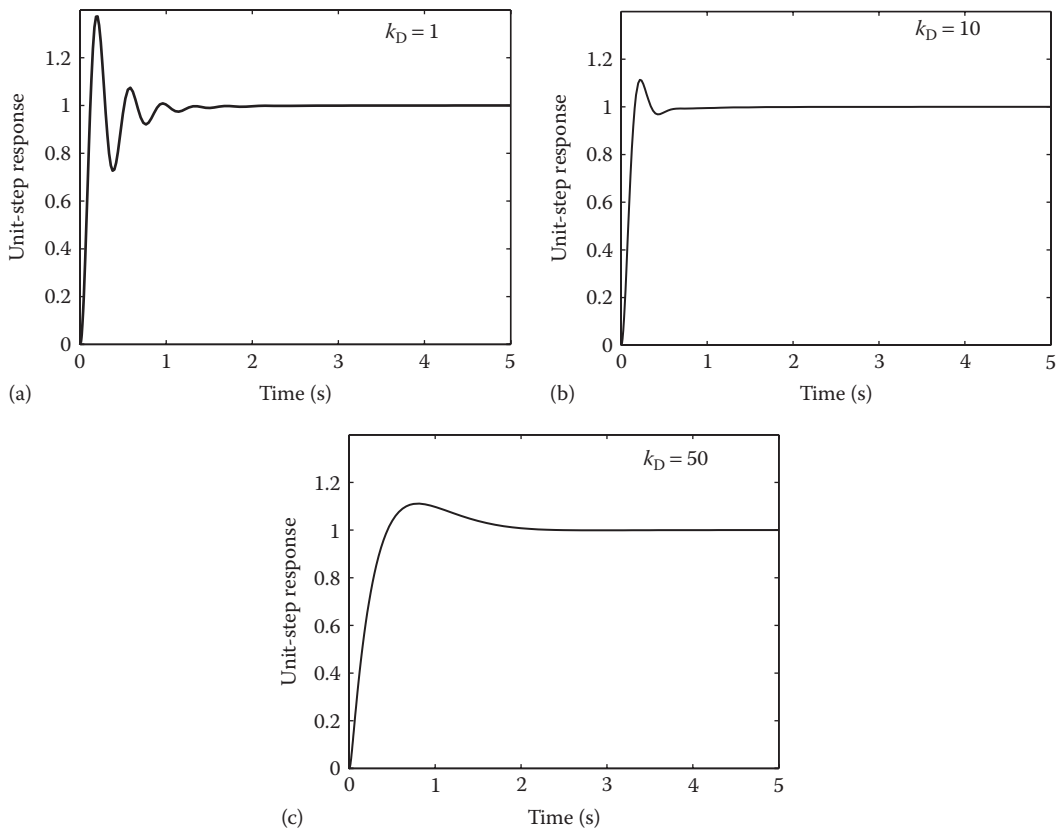
Example 10.12: Proportional–Integral–Derivative Control

Use the Simulink block diagram built in Example 10.10 to find the unit-step responses for $k_D = 1, 10$, and 50 . Set $k_p = 250$ and $k_I = 500$. Discuss the effects of the derivative term on the unit-step response.

Solution

Figure 10.34 shows the unit-step responses of PID feedback control for $k_D = 1, 10$, and 50 . As k_D increases, the overshoot of the unit-step response becomes smaller. This implies that the damping of the feedback control system becomes larger. However, a large derivative control gain k_D leads to a slower response.

In summary, a larger proportional gain k_p results in a faster response and a smaller steady-state error. However, an excessively large proportional gain k_p leads to lightly damped

**FIGURE 10.34**

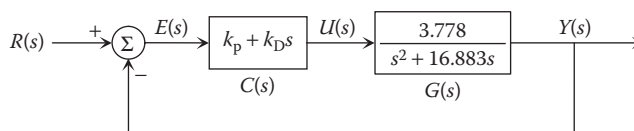
Responses of PID control for (a) $k_D = 1$, (b) $k_D = 10$, and (c) $k_D = 50$.

oscillations and even instability. A larger integral gain k_I eliminates steady-state errors more quickly, but reduces damping and leads to a larger overshoot. A larger derivative control decreases the overshoot, but slows down the speed of the response.

If a PID feedback control system is second-order, then any two free control gains (among k_P , k_I , and k_D) can be determined based on the system stability and performance requirements.

Example 10.13: Proportional-Derivative Control of a Direct Current Motor-Driven Cart

Consider the feedback control system shown in Figure 10.35, in which the plant is the DC motor-driven cart given in Example 10.2. The input to the plant is the voltage applied to

**FIGURE 10.35**

Block diagram of a feedback control system.

the DC motor, and the output is the position of the cart. Design a proportional-derivative (PD) controller such that the maximum overshoot in the response to a unit-step reference input is less than 10%, and the rise time is less than 0.15 s.

Solution

The closed-loop transfer function $Y(s)/R(s)$ is

$$\frac{Y(s)}{R(s)} = \frac{C(s)G(s)}{1 + C(s)G(s)} = \frac{(k_p + k_D s) \frac{3.778}{s^2 + 16.883s}}{1 + (k_p + k_D s) \frac{3.778}{s^2 + 16.883s}} = \frac{3.778k_D s + 3.778k_p}{s^2 + (16.883 + 3.778k_D)s + 3.778k_p}$$

which is a second-order system. The closed-loop characteristic equation is

$$s^2 + (16.883 + 3.778k_D)s + 3.778k_p = 0$$

where the coefficients are related to the natural frequency and the damping ratio of the closed-loop system via

$$16.883 + 3.778k_D = 2\zeta\omega_n$$

$$3.778k_p = \omega_n^2$$

Note that there are two requirements for the transient response of the closed-loop system, that is, $M_p < 10\%$ and $t_r < 0.15$ s. To satisfy these two requirements, a set of reasonable values of ω_n and ζ can be approximated using the relationships given in [Section 10.2](#).

$$M_p = e^{-\pi\zeta/\sqrt{1-\zeta^2}} < 10\%$$

$$t_r \approx \frac{1.12 - 0.078\zeta + 2.230\zeta^2}{\omega_n} < 0.15 \text{ s}$$

The requirement for overshoot indicates

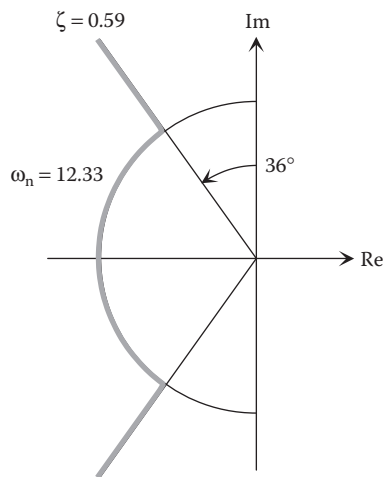
$$\zeta > 0.59$$

Letting $\zeta = 0.59$ and substituting it into the requirement for rise time gives

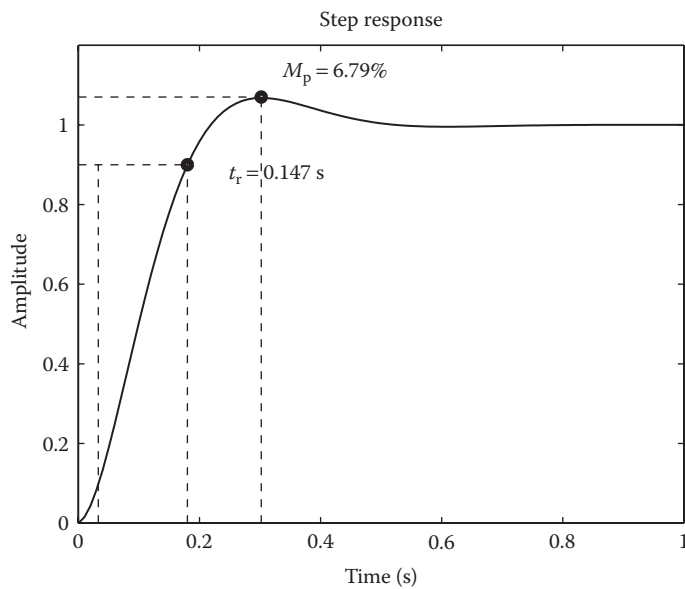
$$\omega_n > 12.33 \text{ rad/s}$$

This region is shown in [Figure 10.36](#), which can be used as a starting point for control design. Note that damping slows the motion of the system. Thus, for a damping ratio higher than 0.59, the critical value of natural frequency should be higher than 12.33 rad/s to speed up the motion of the system. If the closed-loop poles are located to the left of the gray boundary in [Figure 10.36](#), then the closed-loop response to a unit-step reference input will very likely meet the desired requirements.

Choosing $\zeta = 0.65$ and $\omega_n = 13.5$ rad/s yields $k_p = 48.24$ and $k_D = 0.18$. [Figure 10.37](#) shows the response of the closed-loop system to a unit-step reference input.

**FIGURE 10.36**

Allowable region of the closed-loop poles in the s -plane.

**FIGURE 10.37**

Unit-step response of the cart system with PD control $C(s) = 48.24 + 0.18 s$.

10.4.4 Ziegler–Nichols Tuning of PID Controllers

As mentioned earlier, all the methods considered in this chapter are model-based control, which requires that a dynamic model of the plant be available. To avoid this requirement, in the early 1940s, Ziegler and Nichols conducted numerous experiments and proposed two useful tuning methods for determining the PID control gains. The form of the PID controller used by Ziegler and Nichols is given by

$$C(s) = k_p \left(1 + \frac{1}{T_I s} + T_D s \right) \quad (10.41)$$

where:

T_I is the integral time

T_D is the derivative time

The gain parameters T_I and T_D are related to the parameters k_p , k_I , and k_D through $k_I = k_p/T_I$ and $k_D = k_p T_D$.

For the first method, known as the reaction curve method, a step response of the plant is measured. As shown in Figure 10.38, the S-shaped curve is characterized by two constants, lag time L and reaction rate R , which are determined by drawing a line tangent to the curve and finding the intersections of the tangent line with the time axis and the steady-state level line. The values of k_p , T_I , and T_D can be set using the parameters L and R according to the rules in Table 10.1.

For the second method, known as the ultimate sensitivity method, the frequency of the oscillations of the plant at the limit of stability is measured. To use this method, a proportional feedback control is applied to the plant and the proportional gain is increased until the closed-loop system becomes marginally stable. The corresponding proportional gain is defined as K_u , also called the ultimate gain. Figure 10.39 shows the response of a marginally stable system. It is a pure harmonic response, in which the period of oscillation P_u can be measured, also known as the ultimate period. The values of k_p , T_I , and T_D can then be set using the parameters K_u and P_u according to the rules in Table 10.2.

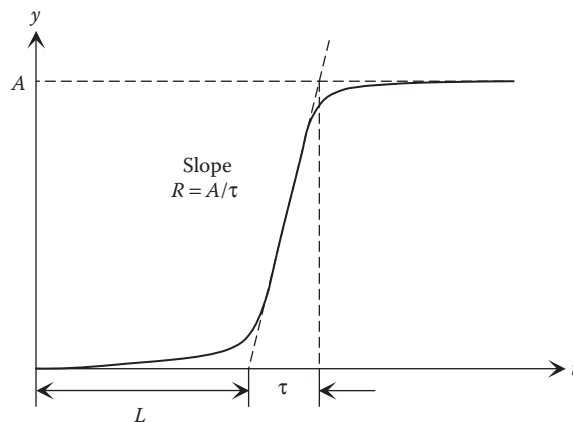


FIGURE 10.38
Reaction curve method.

TABLE 10.1

Ziegler–Nichols Tuning Based on Reaction Curve

Type of Controller	Optimum Gains
P	$k_p = 1/(RL)$
PI	$k_p = 0.9/(RL)$, $T_I = L/0.3$
PID	$k_p = 1.2/(RL)$, $T_I = 2L$, $T_D = 0.5L$

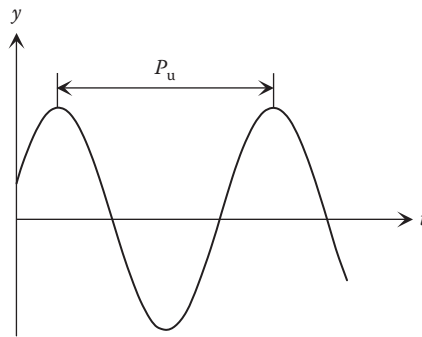


FIGURE 10.39
Response of a marginally stable system.

TABLE 10.2

Ziegler–Nichols Tuning Based on Ultimate Gain and Ultimate Period

Type of Controller	Optimum Gains
P	$k_p = 0.5K_u$
PI	$k_p = 0.45K_u$, $T_I = P_u/1.2$
PID	$k_p = 0.6K_u$, $T_I = P_u/2$, $T_D = P_u/8$

PROBLEM SET 10.4

1. Figure 10.40 shows a negative feedback control system.
 - a. Design a P controller such that the damping ratio of the closed-loop system is 0.5.
 - b. Estimate the rise time, overshoot, and 2% settling time in the unit-step response for the closed-loop system.
2. Consider the negative feedback control system shown in Figure 10.41.
 - a. Design a P controller such that the maximum overshoot in the response to a unit-step reference input is less than 15%, the 2% settling time is less than 1 s, and the rise time is less than 0.2 s.

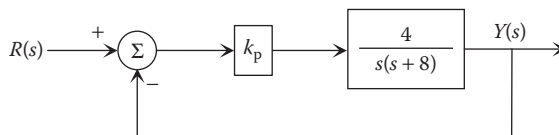


FIGURE 10.40
Problem 1.

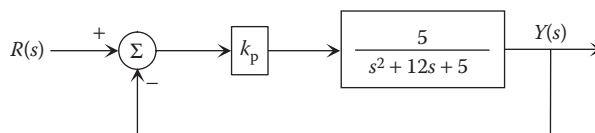


FIGURE 10.41
Problem 2.

- b.  Use MATLAB to plot the unit-step response of the closed-loop system. Find the overshoot, 2% settling time, and rise time. If the time-domain specifications exceed the requirements, fine-tune and reduce them to be approximately the specified values or less.
3. Consider the feedback control system shown in [Figure 10.42](#).
 - a. Design a PD controller such that the closed-loop poles are at $p_{1,2} = -6 \pm 8j$.
 - b.  Use MATLAB to plot the unit-step response of the closed-loop system. Find the rise time, overshoot, peak time, and 1% settling time.
4. Consider the feedback control system shown in [Figure 10.43](#).
 - a. Find the values for k_p and T_D such that the maximum overshoot in the response to a unit-step reference input is less than 5% and the 2% settling time is less than 0.5 s.
 - b. Compute the steady-state error of the closed-loop system to a unit-step reference input.
 - c.  Verify the results of Parts (a) and (b) using MATLAB by plotting the unit-step response of the closed-loop system. If the maximum overshoot and the settling time exceed the requirements, fine-tune and reduce them to be approximately the specified values or less.
5. Consider the feedback control system shown in [Figure 10.44a](#).
 - a. If the desired closed-loop poles are located within the shaded regions shown in [Figure 10.44b](#), determine the corresponding ranges of ω_n and ζ of the closed-loop system.
 - b. Design a PI controller such that the closed-loop poles are at $p_{1,2} = -9 \pm 12j$.
 - c. Compute the steady-state errors of the plant and the closed-loop system to a unit-step reference input.
 - d.  Verify the results in Part (c) using MATLAB by plotting the unit-step responses of the plant and the closed-loop system.

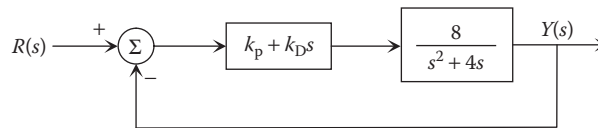


FIGURE 10.42
Problem 3.

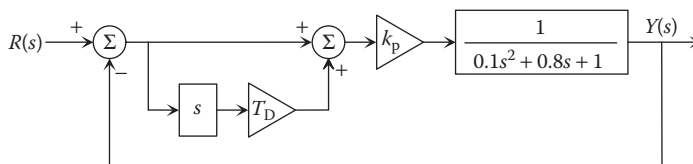


FIGURE 10.43
Problem 4.

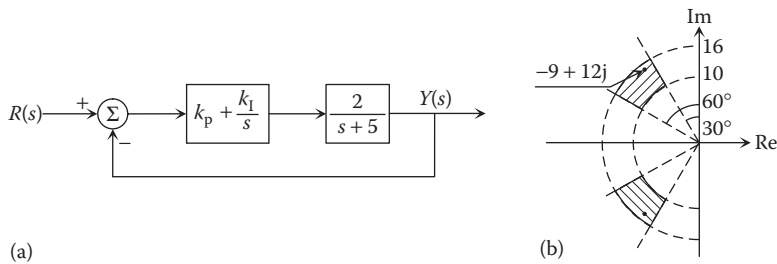


FIGURE 10.44
Problem 5.

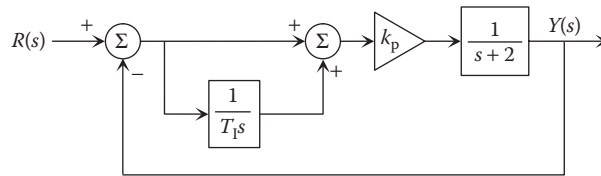


FIGURE 10.45
Problem 6.

6. Consider the feedback control system shown in [Figure 10.45](#).
 - a. Find the values for k_p and T_I such that the maximum overshoot in the response to a unit-step reference input is less than 15% and the peak time is less than 0.4 s.
 - b.  Verify the results in Part (a) using MATLAB by plotting the unit-step response of the closed-loop system. If the maximum overshoot and the peak time exceed the requirements, fine-tune and reduce them to be approximately the specified values or less.
7. The unit-step response of a plant is shown in [Figure 10.46](#).
 - a. The lag time L and the reaction rate R can be determined from the figure. Find the P, PI, and PID controller parameters using the Ziegler–Nichols reaction curve method.

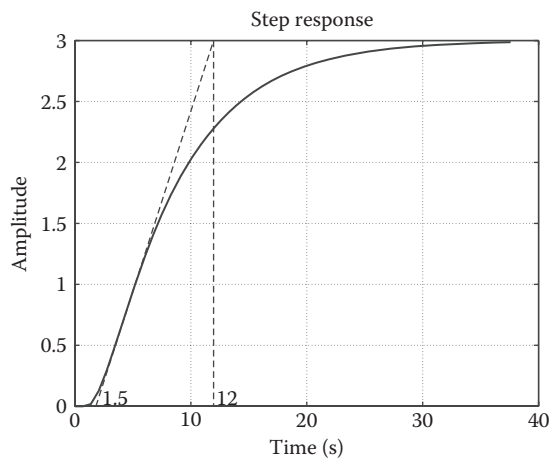


FIGURE 10.46
Problem 7.

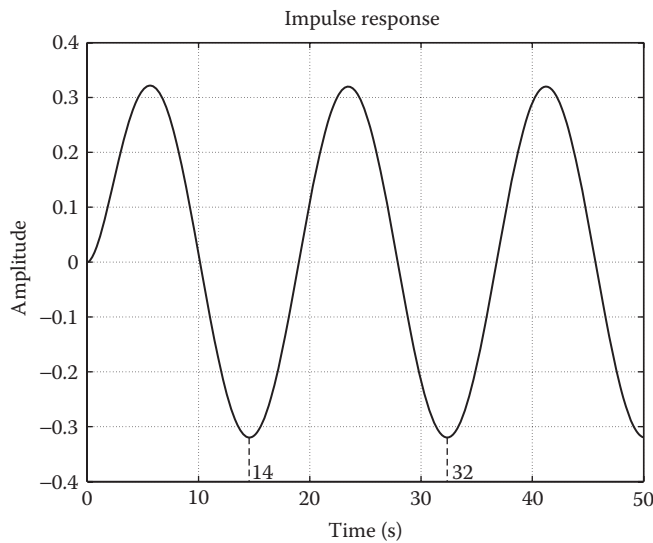


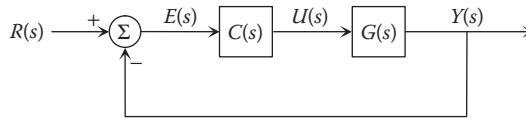
FIGURE 10.47
Problem 8.

- b.  Assume that the transfer function of the plant is $3/(10s^2 + 8s + 1)$. Use MATLAB to plot the unit-step response of the closed-loop system with P, PI, or PID control.
8. Consider a proportional feedback control system. As shown in [Figure 10.47](#), the closed-loop system becomes marginally stable when the proportional gain is 0.75. Find the P, PI, and PID controller parameters using the Ziegler–Nichols ultimate sensitivity method.

10.5 Root Locus

Beginning with this section, and for the remainder of this chapter, the discussion focuses on how to design a feedback controller to meet stability and performance requirements. Root locus, Bode plot, and state-space techniques are three control design methods that are introduced in [Sections 10.5](#) through [10.7](#), respectively. To simplify the discussion, all controllers in those three sections are limited to proportional feedback control.

As presented in [Section 10.2](#), the time-domain specifications, such as rise time t_r , overshoot M_p , peak time t_p , and settling time t_s , are related to the natural frequency ω_n and the damping ratio ζ , both of which can be used to express the pole locations of a second-order system in the s -plane. Thus, the dynamic response of a system can be influenced by changing the system's pole locations. Root locus, developed by W. R. Evans in the late 1940s, is a graphical design technique that shows how changes in one of the system parameters will modify roots of the closed-loop characteristic equation, or the closed-loop poles, and thus change the dynamic response of the system. In this section, we first introduce the procedure to sketch the root locus of a feedback control system. Then, we discuss ways to analyze the

**FIGURE 10.48**

Block diagram of a basic feedback control system.

stability and performance of the closed-loop system based on the root locus. Finally, we learn how to design a proportional feedback controller using the root locus technique.

10.5.1 Root Locus of a Basic Feedback System

Consider a basic feedback control system as shown in Figure 10.48, in which $G(s)$ is the transfer function of the plant. The controller is assumed to be a proportional gain, $C(s) = K$. The closed-loop transfer function is

$$\frac{Y(s)}{R(s)} = \frac{C(s)G(s)}{1 + C(s)G(s)} = \frac{KG(s)}{1 + KG(s)} \quad (10.42)$$

and the characteristic equation is

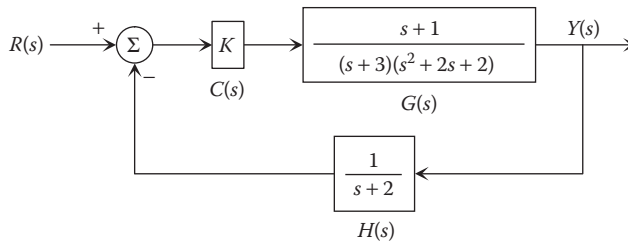
$$1 + KG(s) = 0 \quad (10.43)$$

Note that the roots of the closed-loop characteristic equation are the poles of the closed-loop system. As observed from Equation 10.43, the closed-loop poles are affected by the value of K . When K varies from 0 to ∞ , the closed-loop poles will move around the s -plane and create a trajectory, or a locus of poles.

Intuitively, a root locus can be constructed by changing the value of K from 0 to ∞ , solving the characteristic equation for the roots, and plotting the poles in the s -plane. However, it was difficult to obtain the poles for high-order systems before the availability of mathematical and scientific computing software. In the late 1940s, Evans developed rules to plot a locus, without actually solving for the roots of the characteristic equation. The following example shows the step-by-step procedure to manually construct a root locus.

Example 10.14: Root Locus Sketching

For the system in Figure 10.49, find the locus of closed-loop poles with respect to K .

**FIGURE 10.49**

Block diagram for the feedback control in Example 10.14.

Solution

Step 1: Express the closed-loop characteristic equation in the form of

$$1 + K \frac{b(s)}{a(s)} = 0 \quad (10.44)$$

From [Figure 10.49](#), the closed-loop transfer function is

$$\frac{Y(s)}{R(s)} = \frac{C(s)G(s)}{1 + C(s)G(s)H(s)}$$

Thus, the closed-loop characteristic equation is

$$1 + C(s)G(s)H(s) = 1 + K \frac{s+1}{(s+3)(s^2+2s+2)} \frac{1}{s+2} = 0$$

which is in the form of $1 + Kb(s)/a(s) = 0$ with

$$\begin{aligned} a(s) &= (s+3)(s^2+2s+2)(s+2) \\ b(s) &= s+1 \end{aligned}$$

Denote $L(s) = b(s)/a(s)$, where $L(s)$ is the loop gain. Note that the roots of $a(s)$ are the poles of $L(s)$ and the roots of $b(s)$ are the zeros of $L(s)$. The number of poles determines the number of branches of the root locus.

Step 2: Draw the axes of the s -plane to a suitable scale and mark a cross symbol “x” for each pole of $L(s)$ and a circle symbol “o” for each zero of $L(s)$.

RULE 1: Assume that $L(s)$ has n poles and m zeros. The n branches of the locus start at the poles of $L(s)$ and the m of these branches end at the zeros of $L(s)$.

Solving $a(s) = 0$ for the poles gives

$$p_1 = -3, \quad p_{2,3} = -1 \pm j, \quad p_4 = -2$$

Solving $b(s) = 0$ for the zero gives

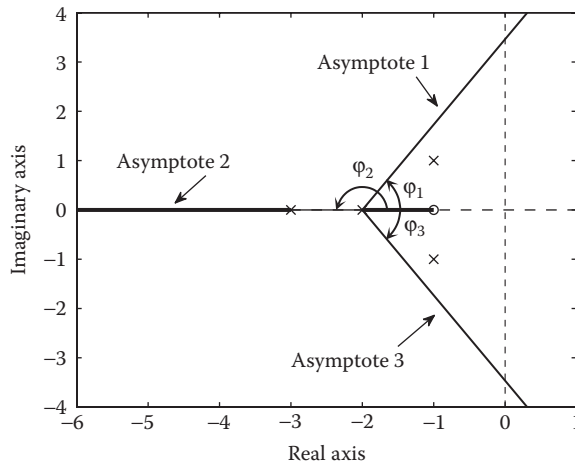
$$z_1 = -1$$

The locations of the four poles and one zero are shown in [Figure 10.50](#).

Step 3: Find the real axis portions of the locus.

RULE 2: The portions of the root locus on the real axis are to the left of an odd number of poles and zeros.

This implies that a point on the real axis is part of the root locus if there is an odd number of poles and zeros to its right. As shown in [Figure 10.50](#), there are two poles, -2 and -3 , and one zero, -1 , located on the real axis, which is divided into four segments, $(-\infty, -3)$, $(-3, -2)$, $(-2, -1)$, and $(-1, +\infty)$. To demonstrate RULE 2, consider a point within $(-2, -1)$. There are two complex poles and one zero (three in total), to its right. Because three is an odd number, this portion of the real axis is part of the root locus. Similarly, the segment $(-\infty, -3)$ is also part of the root locus because there are four poles and one zero (a total of five), to its right. The thick solid lines in [Figure 10.50](#) represent the portions of the root locus on the real axis.

**FIGURE 10.50**

Real axis portions of the root locus and asymptotes for Example 10.14.

Step 4: Draw the asymptotes for large values of K .

RULE 3: For large s and K , $n - m$ branches of the locus are asymptotic to lines radiating out from the point $s = \alpha$ on the real axis at an angle ϕ_l , where

$$\alpha = \frac{\sum \text{Re}(p_i) - \sum \text{Re}(z_i)}{n - m} \quad (10.45)$$

$$\phi_l = \frac{180^\circ + 360^\circ(l - 1)}{n - m}, \quad l = 1, 2, \dots, n - m \quad (10.46)$$

In our case, $n = 4$ and $m = 1$. Thus, there are three asymptotes, which radiate from a centroid α on the real axis

$$\alpha = \frac{(-1 - 1 - 2 - 3) - (-1)}{3} = -2$$

with angles of

$$\phi_1 = \frac{180^\circ + 360^\circ(1 - 1)}{3} = 60^\circ$$

$$\phi_2 = \frac{180^\circ + 360^\circ(2 - 1)}{3} = 180^\circ$$

$$\phi_3 = \frac{180^\circ + 360^\circ(3 - 1)}{3} = 300^\circ \quad \text{or} \quad -60^\circ$$

The centroid and asymptotes are shown in [Figure 10.50](#).

Step 5: Compute the departure and arrival angles.

RULE 4: The angle of departure of a branch of the locus from a pole is given by

$$\phi_{\text{dep}} = \sum \psi_i - \sum \phi_i + 180^\circ \quad (10.47)$$

where:

$\sum \phi_i$ is the sum of the angles from this pole to the remaining poles

$\sum \psi_i$ is the sum of the angles from this pole to all the zeros

The angle of arrival of a branch at a zero is given by

$$\psi_{\text{arr}} = \sum \phi_i - \sum \psi_i + 180^\circ \quad (10.48)$$

where:

$\sum \psi_i$ is the sum of the angles from this zero to the remaining zeros

$\sum \phi_i$ is the sum of the angles from this zero to all the poles

Note that Equations 10.47 and 10.48 are valid for nonrepeated poles and zeros. The formula used for computing the departure or arrival angles from a repeated pole or to a repeated zero can be found in control books.

In [Figure 10.50](#), two branches of the locus on the real axis have been completely drawn. One of them starts from the pole at -2 and ends at the zero at -1 , and the other starts from the pole at -3 and ends at $-\infty$ by approaching the second asymptote. There is no need to compute the departure or arrival angles for the poles and the zero on the real axis. Note that there is a pair of complex conjugate poles at $-1 \pm j$, from each of which one branch of the locus starts. Selecting the pole at $-1+j$ and applying Equation 10.47 gives

$$\phi_{\text{dep}} = \sum \psi_i - \sum \phi_i - 180^\circ = \psi_1 - (\phi_1 + \phi_2 + \phi_3) - 180^\circ$$

As sketched in [Figure 10.51](#), $\psi_1 = 90^\circ$, which is the angle of the line connecting the complex pole at $-1+j$ and the zero with respect to the positive real axis. Similarly, we can determine the angles ϕ_i , as $\phi_1 = 90^\circ$, $\phi_2 = 45^\circ$, and $\phi_3 = \tan^{-1}(\frac{1}{2}) = 26.57^\circ$. Thus,

$$\phi_{\text{dep}} = 90^\circ - (90^\circ + 45^\circ + 26.57^\circ) - 180^\circ = -251.57^\circ = 108.43^\circ$$

The departure angle from the pole at $-1-j$ is -108.43° because the root locus is symmetric about the real axis.

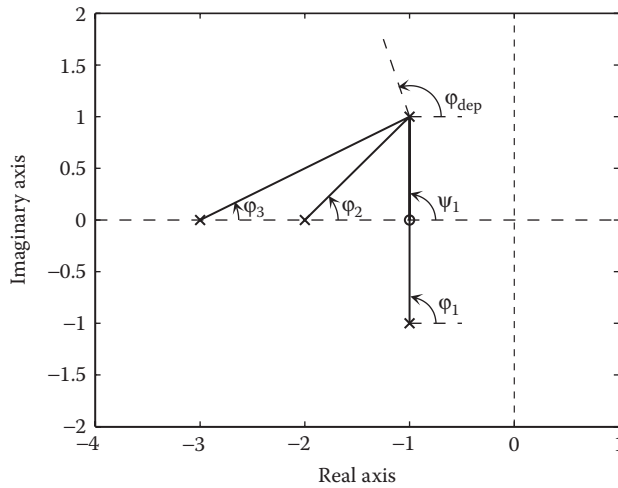
Step 6: Determine the points where the root locus crosses the imaginary axis.

RULE 5: The root locus crosses the imaginary axis at points where the characteristic equation satisfies

$$1 + K \frac{b(j\omega)}{a(j\omega)} = 0 \quad (10.49)$$

or

$$a(j\omega) + Kb(j\omega) = 0 \quad (10.50)$$

**FIGURE 10.51**

Departure angles for Example 10.14.

Note that if a point on the imaginary axis is part of the root locus, then it must be a root of the characteristic equation given by Equation 10.44. The point on the imaginary axis can be expressed as $s = j\omega$ and substituting it into Equation 10.44 leads to Equation 10.49 or 10.50. Solving either of them yields the $j\omega$ -axis crossing points and the corresponding values of K .

In this example, we have

$$(s + 3)(s^2 + 2s + 2)(s + 2) + K(s + 1) = 0$$

Substituting $s = j\omega$ and rearranging the equation gives

$$(j\omega)^4 + 7(j\omega)^3 + 18(j\omega)^2 + 22j\omega + 12 + Kj\omega + K = 0$$

Separating the real and imaginary terms results in

$$(\omega^4 - 18\omega^2 + 12 + K) + j(-7\omega^3 + 22\omega + K\omega) = 0$$

which is equivalent to two equations

$$\begin{cases} \omega^4 - 18\omega^2 + 12 + K = 0 \\ -7\omega^3 + 22\omega + K\omega = 0 \end{cases}$$

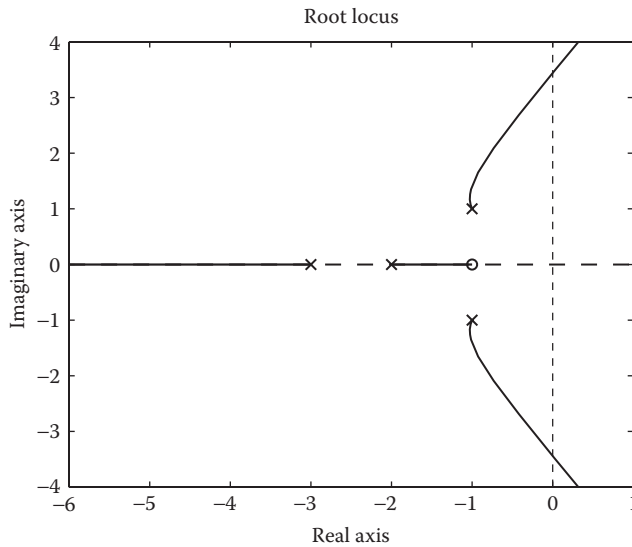
Solving for ω , we obtain

$$\omega_1 = 0, \quad \omega_{2,3} = \pm 3.44, \quad \omega_{4,5} = \pm 0.92j$$

where the last two roots are not valid because ω represents frequency and is real. The corresponding values of K are

$$K_1 = -12, \quad K_{2,3} = 60.91$$

where the first value is not valid because the gain K varies from 0 to $+\infty$. Thus, the root locus crosses the $j\omega$ -axis at $\pm 3.44j$ when $K = 60.91$.

**FIGURE 10.52**

Root locus for Example 10.14.

Step 7: Complete the sketch.

As shown in [Figure 10.52](#), two complex branches of the locus depart from the complex poles, cross the imaginary axis, enter the right-half s -plane, and end at infinity by approaching the asymptotes. Combining the complex branches with the real axis portions of the locus, we have the sketch completed.

Rules 1 through 5 can be used to roughly sketch a root locus. One more rule is available for computing the so-called break-in or break-away points, but this will not be covered in this text. With the availability of MATLAB, these rules are not necessary to plot a root locus. However, learning these rules can help understand classical control design techniques and evaluate the correctness of a computer-generated root locus.

The MATLAB command used to sketch a root locus is `rlocus`, which calls the loop gain $L(s)$. The following is the MATLAB session to generate the root locus shown in [Figure 10.52](#).

```
>> num = [1 1];
>> den = conv(conv([1 3], [1 2 2]), [1 2]);
>> sysL = tf(num, den);
>> rlocus(sysL);
```

10.5.2 Analysis Using Root Locus

Using the root locus technique, it is very easy to determine the stability of a closed-loop system when the proportional gain K varies from 0 to ∞ . For a particular value of K , if any of the poles are in the right-half s -plane, then the corresponding closed-loop system is unstable. If all the poles are in the left-half s -plane, then the closed-loop system is stable. If any of the poles are on the imaginary axis and they are nonrepeated, then the closed-loop system is marginally stable.

From the root locus, we can also obtain information about the performance of a closed-loop system using the concept of a dominant pole. For a system with multiple poles, the pole closest to the origin is called the dominant pole. If the dominant poles are a pair of complex conjugates, the distance between either one of them and the origin is associated with the natural frequency if the system is approximated as a second-order system. If the dominant pole is a real pole, the distance between the pole and the origin is associated with the time constant if the system is approximated as a first-order system. Both the natural frequency and the time constant determine the speed of transient response. *The dominant pole has the slowest speed of response*, and it dominates the effect of all other poles with higher frequencies or lower time constants.

Example 10.15: Analysis Using Root Locus

Refer to the root locus obtained in Example 10.14. Comment on the stability and performance of the closed-loop system when K varies from 0 to ∞ .

Solution

When $K = 0$, which corresponds to having no control, the root locus starts from the poles of the loop gain $L(s)$. As shown in Figure 10.53, all four open-loop poles are located in the left-half s -plane. This implies that the open-loop system is stable. As K increases, the four closed-loop poles move along four different branches of the root locus. For $0 \leq K < 60.91$, all the poles are in the left-half s -plane, and thus, the closed-loop system is stable. When $K = 60.91$, two complex poles, $\pm 3.44j$, appear on the imaginary axis, and thus, the closed-loop system becomes marginally stable. For $K > 60.91$, the two complex branches of the root locus cross the imaginary axis and enter the right-half s -plane. The closed-loop system becomes unstable.

When K varies between 0 and 60.91, the pair of complex poles has smaller absolute value of real part than the other poles. Consequently, they dominate the effect of all other poles, and thus, the closed-loop system exhibits undamped oscillations. For example, the closed-loop poles are -3.29 , -1.66 , and $-1.03 \pm 1.15j$ when $K = 1$. Figure 10.54 is the corresponding closed-loop unit-step response, which has an overshoot of 21.1%.

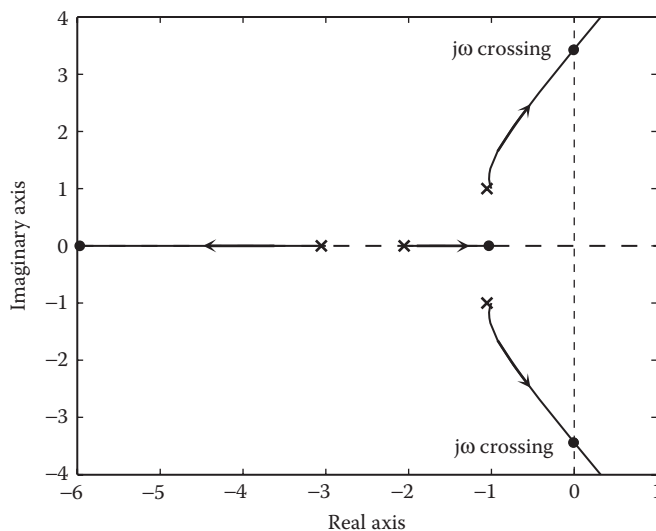
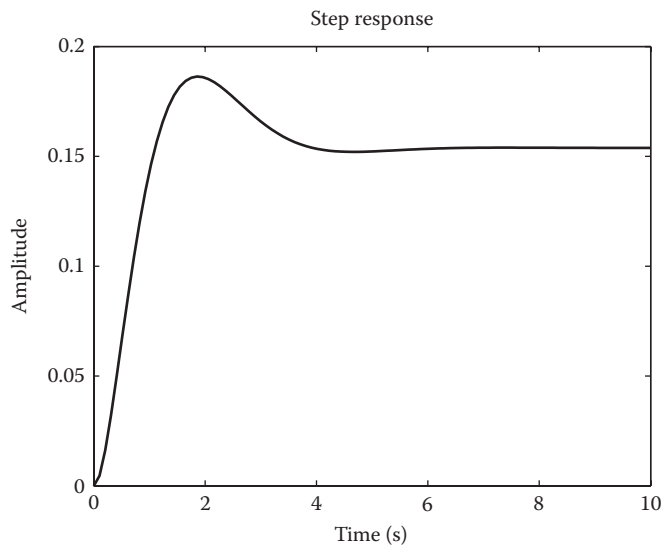


FIGURE 10.53

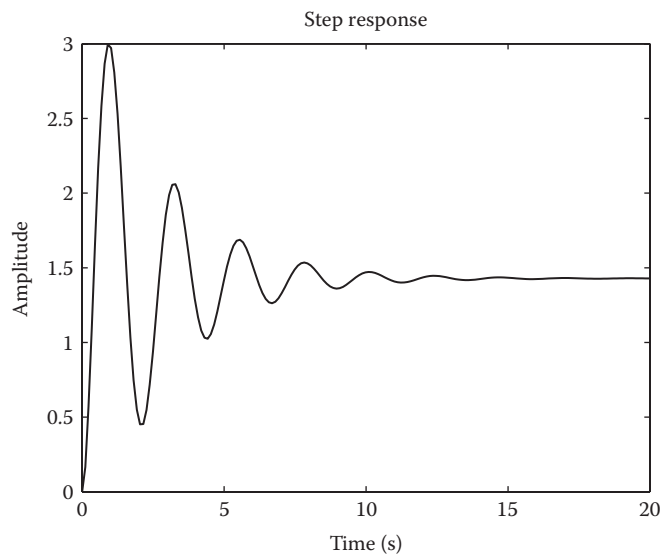
Root locus for Example 10.14 with poles highlighted when $K = 60.91$.

**FIGURE 10.54**

Unit-step response of the closed-loop system when $K = 1$.

As K increases, the closed-loop system will exhibit severe oscillations because the dominant complex poles move toward the imaginary axis and the damping decreases. For example, the closed-loop poles are -5.16 , -1.06 , and $-0.39 \pm 2.74j$ when $K = 30$. [Figure 10.55](#) is the corresponding closed-loop unit-step response, in which the overshoot is as high as 110%.

Note that although the root locus is constructed based on the loop gain $L(s)$, it also gives information on the stability and performance of the closed-loop system varying with respect to the proportional gain K . This is what makes the root locus technique attractive.

**FIGURE 10.55**

Unit-step response of the closed-loop system when $K = 30$.

10.5.3 Control Design Using Root Locus

As we learned in [subsection 10.5.1](#), the root locus is a plot of all possible roots of the closed-loop characteristic equation $1 + KL(s) = 0$ for real positive values of K , which is generally the gain of a proportional controller. It is very easy to select K from a particular root locus, so that the closed-loop system meets the performance specifications.

Example 10.16: Proportional Control Design Using Root Locus

Design a proportional controller for the cart system in Example 10.13 using the root locus technique.

Solution

For proportional feedback control of the DC motor-driven cart, the loop gain $L(s)$ is equal to the transfer function of the plant $G(s)$,

$$L(s) = G(s) = \frac{3.778}{s^2 + 16.883s}$$

The root locus is plotted in [Figure 10.56](#) using the MATLAB command `rlocus`. Note that the closed-loop system is a second-order system. Two closed-loop poles that are real

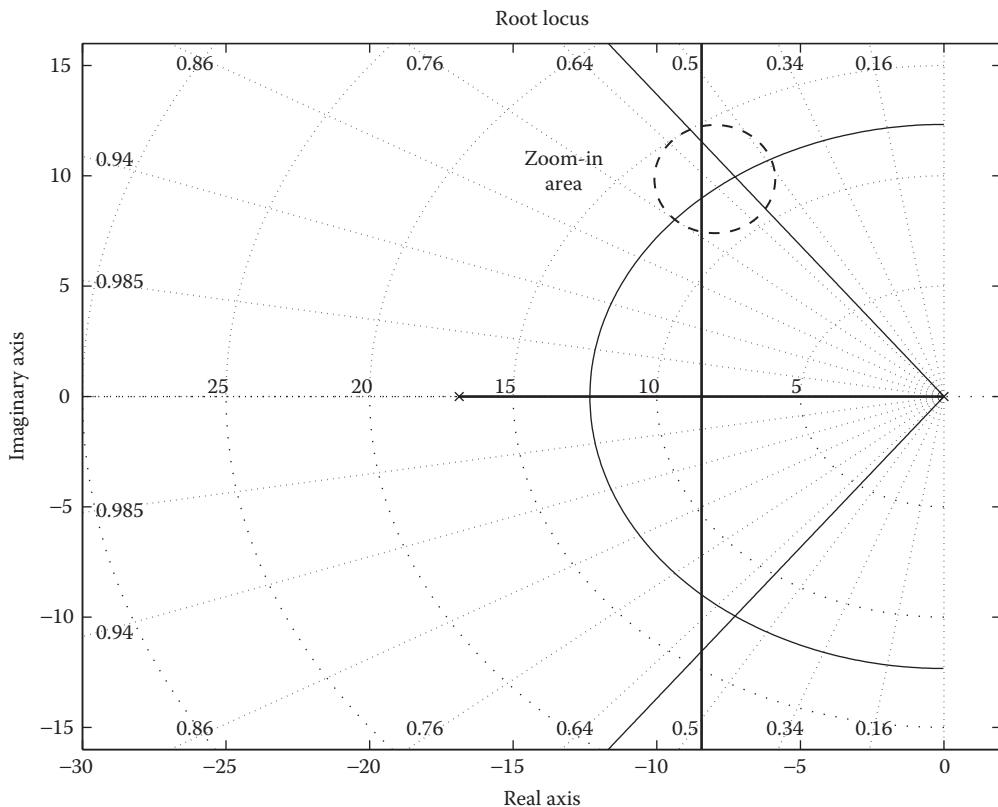


FIGURE 10.56

Root locus for Example 10.16 with grid lines.

appear for small values of K , specifically $K \leq 18.9$. When $K > 18.9$, the closed-loop system has a pair of complex conjugate poles, which move along the complex portion of the root locus as K varies between 18.9 and ∞ .

To select the value of the proportional control gain that will meet the performance specifications, we can turn on the grid lines to the root locus using the following command.

```
>> grid on
```

As discussed in [subsection 10.2.2](#), the semicircles in [Figure 10.56](#) indicate lines of constant natural frequencies ω_n and the diagonal lines indicate constant damping ratios ζ . In this example, we need an overshoot that is less than 10% (which implies $\zeta > 0.59$) and a rise time that is less than 0.15 s (which implies $\omega_n > 12.33$ rad/s). In [Figure 10.56](#), the solid diagonal lines indicate pole locations with a damping ratio of about 0.59. In between these lines, the poles have $\zeta > 0.59$ and outside of the lines $\zeta < 0.59$. The solid semicircle is the locus of all poles with $\omega_n = 12.33$ rad/s, whereas those inside the semicircle have $\omega_n < 12.33$ rad/s, and those outside correspond to $\omega_n > 12.33$ rad/s. Thus, only the part of the root locus between the two diagonal lines and outside of the semicircle is acceptable.

[Figure 10.57](#) zooms in on the desired region, in which the vertical line is a part of the root locus. Left-clicking the root locus, you will see the values of the pole and the gain that is required to place one of the closed-loop poles at that particular location. Holding

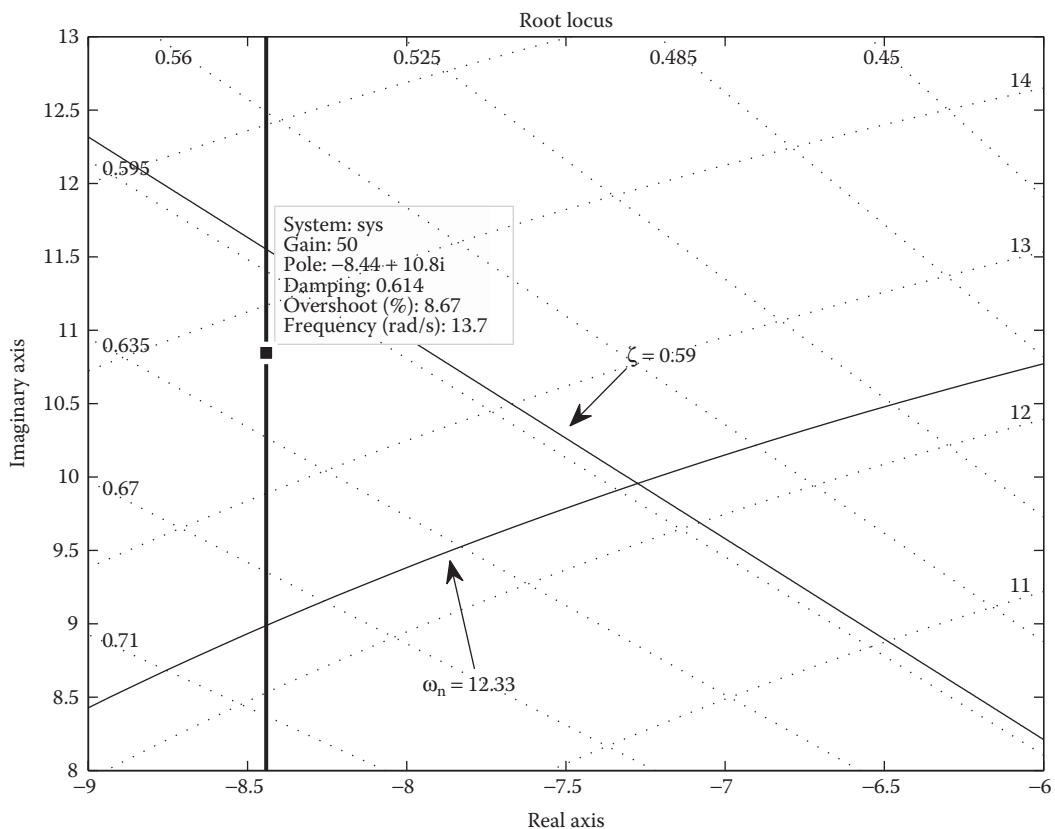
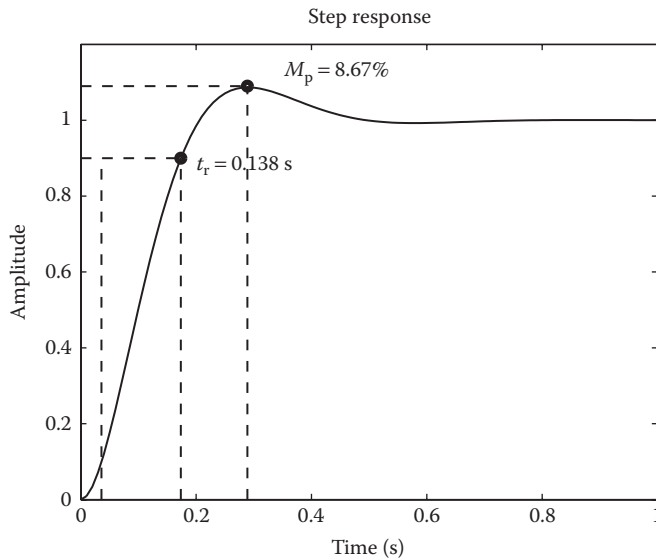


FIGURE 10.57

Zoom-in of the designed region in Example 10.16.

**FIGURE 10.58**

Unit-step response of the cart system with proportional feedback control, $K = 50$.

down the left mouse button and moving the mouse along the root locus, you can see the values of the pole and the gain varying correspondingly.

Let us select $K = 50$. Figure 10.58 is the corresponding unit-step response of the closed-loop system with proportional feedback control. The closed-loop system meets the given specifications.

Note that the closed-loop poles cannot be placed arbitrarily in the s -plane with only a static proportional controller since the shape of the root locus is fixed for a given plant. A more useful design can be obtained by adding a pole or zero to the controller and making it a dynamic controller. This results in so-called lead or lag compensators, $C(s) = (s + z)/(s + p)$. The reader can refer to the control systems books for more details.

PROBLEM SET 10.5

1. Roughly sketch the root locus with respect to K for the equation of $1 + KL(s) = 0$ and the following choices for $L(s)$. Make sure to give the asymptotes, arrival or departure angles, and points crossing the imaginary axis. Verify your results using MATLAB.

- a. $L(s) = \frac{1}{(s+1)(s+3)}$
- b. $L(s) = \frac{1}{(s+1)(s+3)(s+7)}$
- c. $L(s) = \frac{s+5}{(s+1)(s+3)(s+7)}$
- d. $L(s) = \frac{s(s+5)}{(s+1)(s+3)(s+7)}$

2. Repeat Problem 1 for the following choices for $L(s)$.
- $L(s) = \frac{1}{s^2 + 2s + 10}$
 - $L(s) = \frac{1}{(s+4)(s^2 + 2s + 10)}$
 - $L(s) = \frac{s^2 + 4s + 8}{(s+4)(s^2 + 2s + 10)}$
 - $L(s) = \frac{(s+1)(s^2 + 4s + 8)}{(s+4)(s^2 + 2s + 10)}$
3. A control system is represented using the block diagram shown in Figure 10.59. Sketch the root locus with respect to the proportional control gain K . Determine all the values of K for which the closed-loop system is stable.
4.  A control system is represented using the block diagram shown in Figure 10.60, in which the parameter a is subjected to variations. Sketch the root locus with respect to the parameter a . Determine all the values of a for which the closed-loop system is stable.
5.  Figure 10.61 shows the root locus of a unity negative feedback control system, where K is the proportional control gain.
- Determine the transfer function of the plant. Use MATLAB to plot the root locus based on your choice of the plant, compare it with the root locus shown in Figure 10.61, and check the accuracy of your plant transfer function.
 - Give comments on the stability of the closed-loop system when K varies from 0 to ∞ .
 - Give comments on the transient performance of the closed-loop system when $K = 0.5$ and 5. Use MATLAB to plot the corresponding unit-step responses and verify your comments.

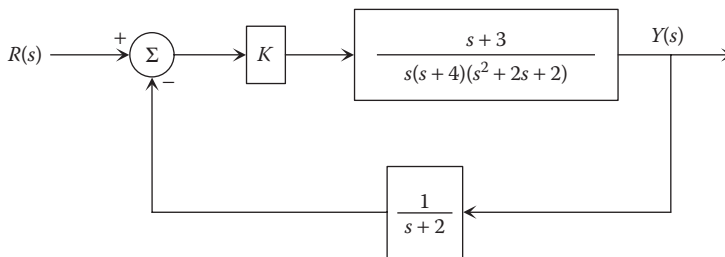


FIGURE 10.59
Problem 3.

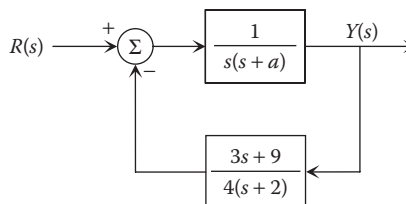
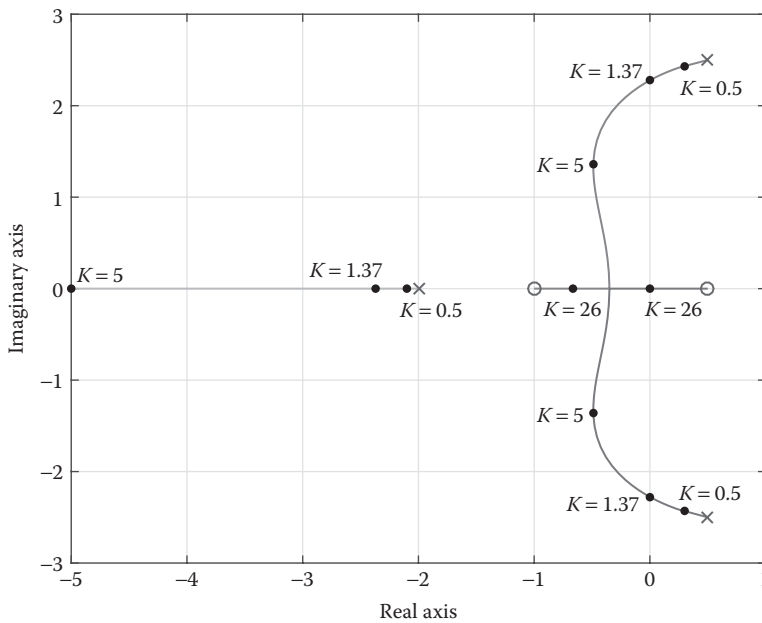


FIGURE 10.60
Problem 4.

**FIGURE 10.61**

Problem 5.

6. Figure 10.62 shows the root locus of a unity negative feedback control system, where K is the proportional control gain.
 - a. Determine the transfer function of the plant. Use MATLAB to plot the root locus based on your choice of the plant, compare it with the root locus shown in Figure 10.62, and check the accuracy of your plant transfer function.
 - b. Find the range of values of K for which the system has damped oscillatory response. What is the value of K when pure harmonic oscillations occur? Also, what is the frequency of pure harmonic oscillations? Use MATLAB to plot the corresponding unit-step response and verify the accuracy of your computed frequency.
7. Consider the feedback system shown in Figure 10.63.
 - a. Find the locus of the closed-loop poles with respect to K .
 - b. Find a value of K such that the maximum overshoot in the response to a unit-step reference input is less than 5%. What is the corresponding steady-state error of the closed-loop system?
 - c. Plot the unit-step response of the closed-loop system to verify the result for Part (b).
8. Consider the feedback system shown in Figure 10.64.
 - a. Find the locus of the closed-loop poles with respect to K .
 - b. Find a value of K such that the maximum overshoot in the response to a unit-step reference input is less than 20% and the 2% settling time is less than 1.1 s.
 - c. Plot the unit-step response of the closed-loop system to verify the result in Part (b).

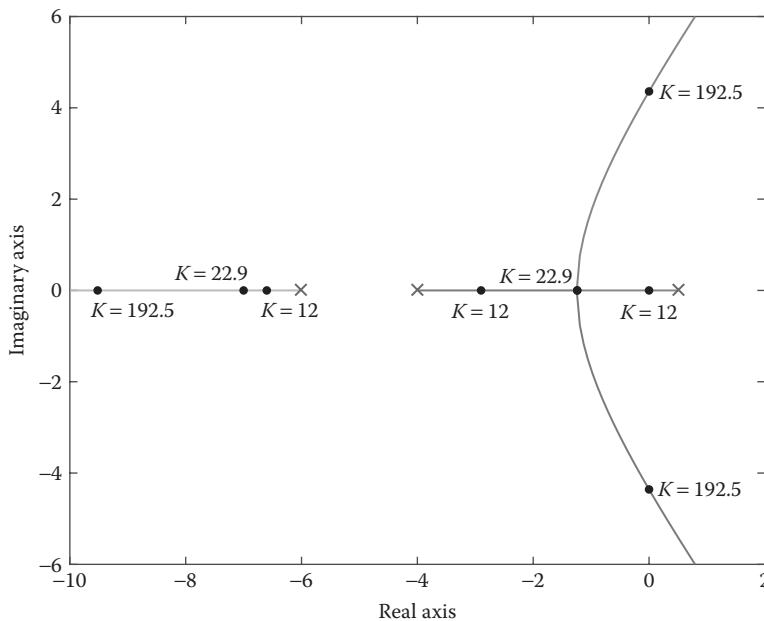


FIGURE 10.62
Problem 6.

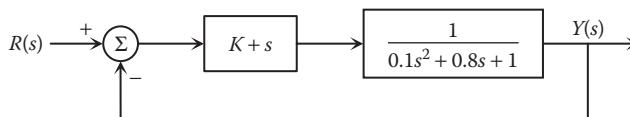


FIGURE 10.63
Problem 7.

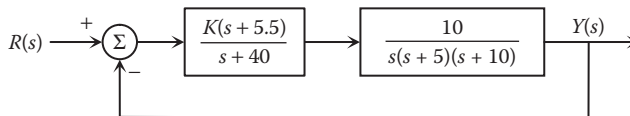


FIGURE 10.64
Problem 8.

10.6 Bode Plot

The Bode plot technique is widely used to display a frequency response function. It also gives useful information for analyzing and designing control systems. Stability criteria can be interpreted using the Bode plot and numerous control design techniques are based on the Bode plot. In [Section 8.4](#), we introduced the concept of the Bode plot and presented the Bode plot of the frequency response function for two fundamental systems, first-order and second-order. In this section, we first discuss how to use the Bode plot to display the frequency response function for a general dynamic system. Then, we learn how the Bode

plot is utilized to determine stability. Finally, we will see how the Bode plot technique is used to design a proportional feedback controller.

10.6.1 Bode Plot of a Basic Feedback System

Consider the basic feedback control system shown in [Figure 10.65](#). The open-loop transfer function is $KG(s)$, which can be written in the form

$$KG(s) = K \frac{(s - z_1)(s - z_2) \cdots (s - z_m)}{(s - p_1)(s - p_2) \cdots (s - p_n)} \quad (10.51)$$

explicitly showing the poles and zeros. Replacing s with $j\omega$ yields the frequency response function

$$KG(j\omega) = K \frac{(j\omega - z_1)(j\omega - z_2) \cdots (j\omega - z_m)}{(j\omega - p_1)(j\omega - p_2) \cdots (j\omega - p_n)} \quad (10.52)$$

The frequency response function can be displayed using two curves: the Bode magnitude plot and the Bode phase plot. By definition, the magnitude of $KG(j\omega)$ in decibels is

$$\begin{aligned} |KG(j\omega)|_{\text{dB}} &= 20 \log_{10} |KG(j\omega)| \\ &= 20 \log_{10} |K| + 20 \log_{10} |j\omega - z_1| + 20 \log_{10} |j\omega - z_2| + \cdots \\ &\quad - 20 \log_{10} |j\omega - p_1| - 20 \log_{10} |j\omega - p_2| - \cdots \end{aligned} \quad (10.53)$$

and the phase of $KG(j\omega)$ is

$$\angle[KG(j\omega)] = \angle K + \angle(j\omega - z_1) + \angle(j\omega - z_2) + \cdots - \angle(j\omega - p_1) - \angle(j\omega - p_2) - \cdots \quad (10.54)$$

Equations 10.53 and 10.54 show that the magnitude (in decibels) and phase of the frequency response function each is the sum of the magnitudes and phases of simple terms, which are similar to each other. If we know how to draw the Bode plot for each individual term, then the composite curve can be obtained by combining all the terms involved.

Depending on the locations of poles or zeros, there are four classes of basic terms:

1. Constant terms K (no pole or zero)
2. Integral or derivative terms $(j\omega)^{\pm n}$ (with pole(s) or zero(s) at the origin)
3. First-order terms $(j\omega\tau + 1)^{\pm 1}$ (with a real pole or zero at $-1/\tau$)
4. Second-order terms $[(j\omega/\omega_n)^2 + 2\zeta j\omega/\omega_n + 1]^{\pm 1}$ (with a pair of complex conjugate poles or zeros at $-\zeta\omega_n \pm j\omega_n\sqrt{1-\zeta^2}$)

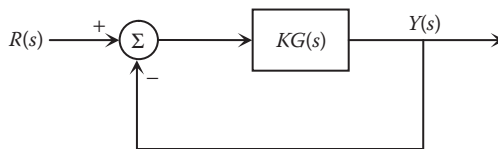


FIGURE 10.65

Simplified feedback control system.

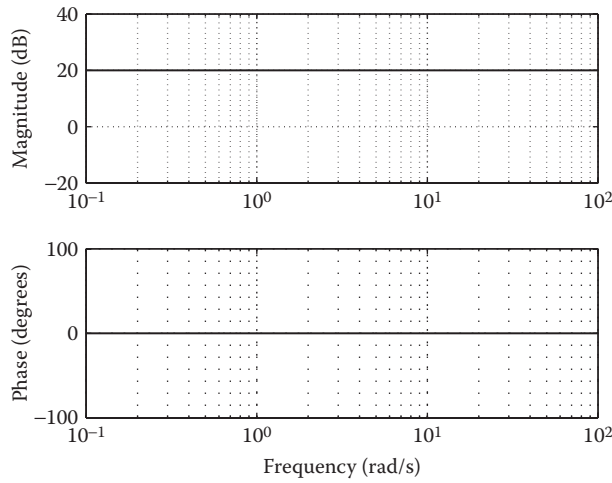


FIGURE 10.66
Bode plot for $K = 10$.

Figure 10.66 is an example of a plot with a constant term $K = 10$. Because K is independent of the frequency, both magnitude and phase in the entire frequency region are horizontal lines. The magnitude in decibels is

$$|K|_{\text{dB}} = 20 \log_{10} |K| \quad (10.55)$$

and the phase is

$$\angle K = 0^\circ \quad (10.56)$$

For integral or derivative terms, the magnitude of $(j\omega)^{\pm n}$ in decibels is

$$|(j\omega)^{\pm n}|_{\text{dB}} = 20 \log_{10} |(j\omega)^{\pm n}| = \pm n \times 20 \log_{10} |j\omega| = \pm 20n \log_{10} \omega \quad (10.57)$$

Note that the magnitude plot is drawn using the logarithmic scale for the frequency, that is, $\log_{10} \omega$. Thus, the magnitude plot of an integral or derivative term is a straight line with a slope $\pm 20n$ dB/decade, which means that the magnitude will change by $\pm 20n$ dB as the frequency increases by a factor of 10. Geometrically, a straight line is uniquely determined by its slope and one point that it goes through. By Equation 10.57, this line always crosses (1 rad/s, 0 dB) regardless of the value of n . Figure 10.67 shows Bode plots for an integral term $1/j\omega$ and a derivative term $j\omega$. Their magnitude plots intersect at (1 rad/s, 0 dB) and the slope is -20 dB/decade for $1/j\omega$ and $+20$ dB/decade for $j\omega$. The phase of $(j\omega)^{\pm n}$ is

$$\angle(j\omega)^{\pm n} = \pm n \angle(j\omega) = \pm n \times 90^\circ \quad (10.58)$$

which is independent of frequency. As shown in Figure 10.67, the phase is a horizontal line and it is -90° for $1/j\omega$ and $+90^\circ$ for $j\omega$.

The Bode plots of the frequency response function for a first-order and second-order system are shown in Figures 8.23 and 8.25. In this section, we discuss the plotting of asymptotes for more general cases as indicated in the third and fourth classes.

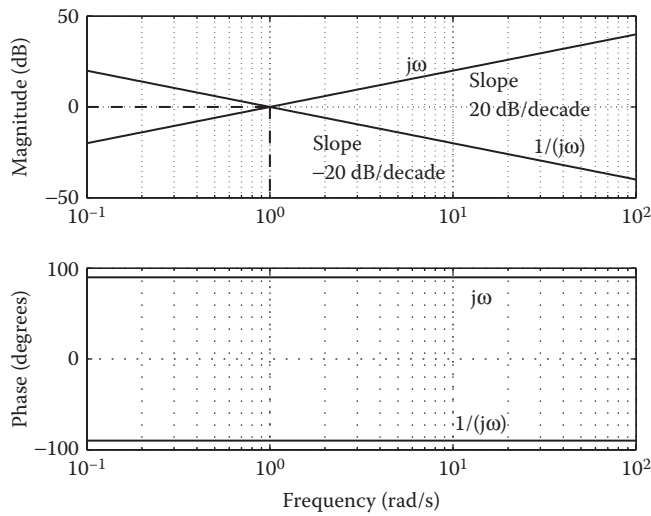


FIGURE 10.67
Bode plot for $1/j\omega$ and $j\omega$.

For the first-order terms $(j\omega\tau + 1)^{\pm 1}$, [Figure 10.68](#) shows the magnitude plots with the asymptotes. Let us take $j\omega\tau + 1$ as an example. At low frequencies, $\omega\tau \ll 1$, we have $j\omega\tau + 1 \approx 1$. Thus, the magnitude approaches a horizontal line crossing 0 dB. At high frequencies, $\omega\tau \gg 1$, we have $j\omega\tau + 1 \approx j\omega\tau$, for which the magnitude is

$$20 \log_{10} |j\omega\tau| = 20 \log_{10}(\omega\tau) = 20 \log_{10} \omega + 20 \log_{10} \tau \approx 20 \log_{10} \omega \quad (10.59)$$

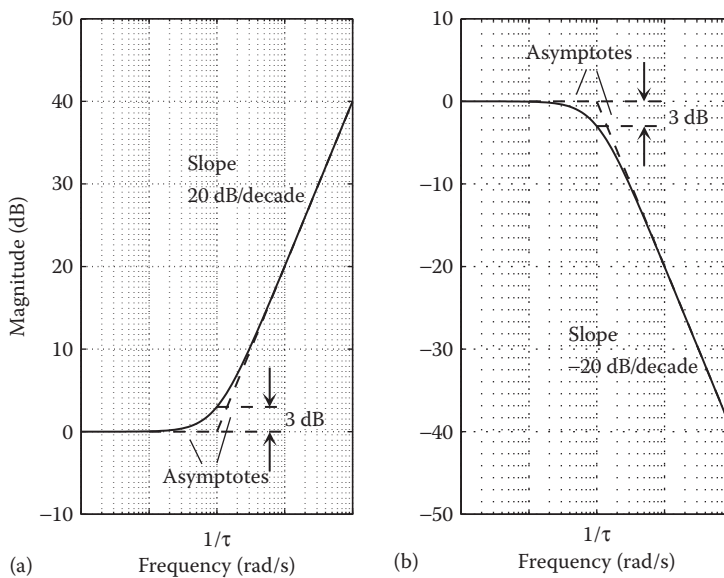


FIGURE 10.68
Magnitude plots for (a) $(j\omega\tau + 1)^{+1}$ and (b) $(j\omega\tau + 1)^{-1}$.

Note that τ is a finite number for a given frequency response function, and its effect on the magnitude can be ignored at very high frequencies. Thus, the magnitude approaches a straight line with a slope of 20 dB/decade. When $\omega\tau = 1$ (or $\omega = 1/\tau$), the magnitude is

$$20 \log_{10} |j + 1| = 20 \log_{10} \sqrt{2} = 3 \text{ dB} \quad (10.60)$$

The corresponding frequency $1/\tau$ is the corner frequency, at which the slope of the asymptote changes from 0 to 20 dB/decade. The magnitude plot in the entire frequency region can be obtained by drawing a smooth curve following the asymptotes with 3 dB above the line at the corner frequency.

Figure 10.69 shows the asymptotes of the phase plots for the terms $(j\omega\tau + 1)^{\pm 1}$. Again, let us take $j\omega\tau + 1$ as an example. The phase can be approximated as $\angle 1 = 0^\circ$ at low frequencies and $\angle(j\omega\tau) = 90^\circ$ at high frequencies. The phase at the corner frequency $1/\tau$ is $\angle(j + 1) = 45^\circ$.

The Bode plot of the second-order terms $[(j\omega/\omega_n)^2 + 2\zeta j\omega/\omega_n + 1]^{\pm 1}$ can be drawn in a similar manner as the first-order terms. The asymptotes for magnitude and phase plots are shown in Figures 10.70 and 10.71, respectively. Unlike the first-order terms, the corner frequency is $\omega = \omega_n$, at which the magnitude changes slope from 0 to +40 dB/decade if the term is in the numerator or to -40 dB/decade if the term is in the denominator. Correspondingly, the phase changes from 0° to $+180^\circ$ or -180° . The magnitude at the corner frequency ω_n greatly depends on the damping ratio ζ . A rough sketch can be made by noting whether the peak is either $20 \log_{10}(1/(2\zeta))$ below or above the asymptotes.

For a general dynamic system, the frequency response function can always be written as the product of several basic terms. Equations 10.53 and 10.54 suggest that the composite magnitude curve and the phase curve are the sum of their respective individual curves. The following example shows how to obtain a quick sketch of the composite curve using the asymptotes of the basic terms.

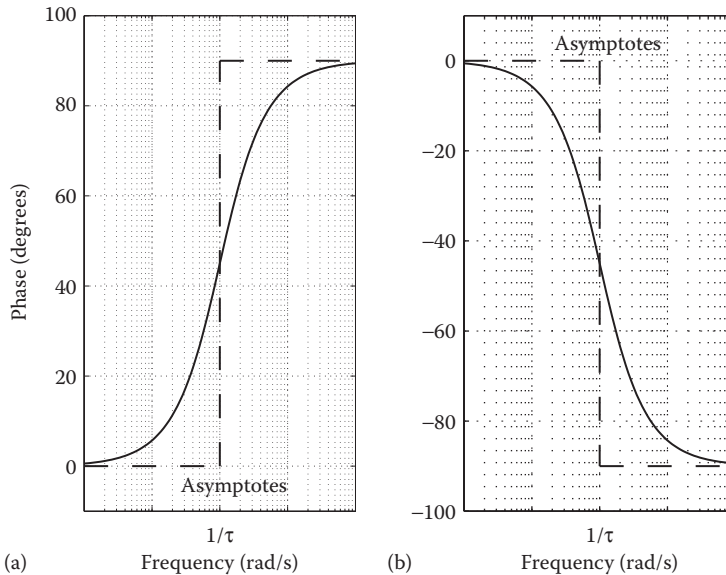
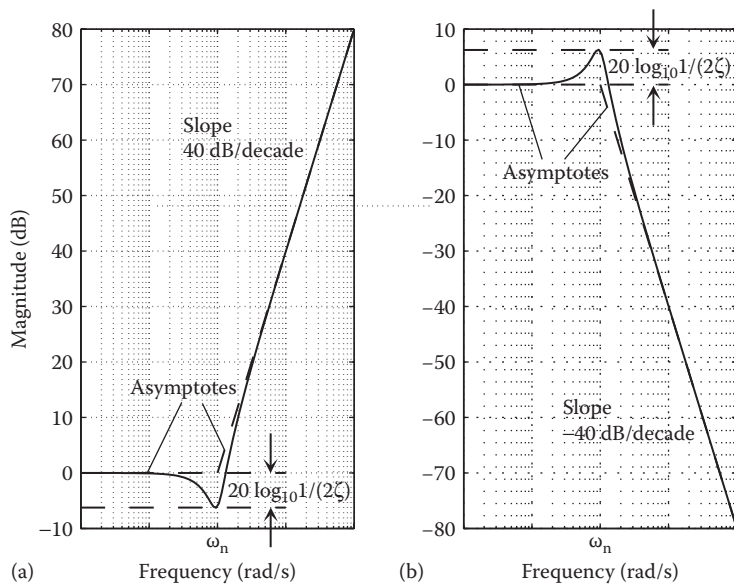
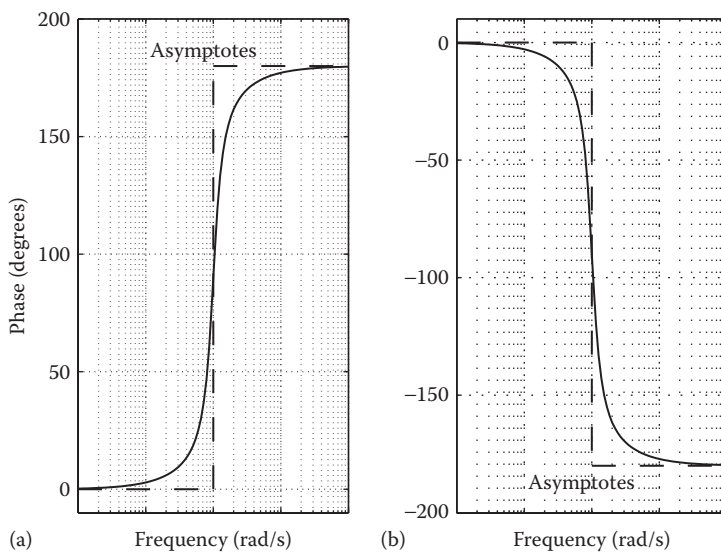


FIGURE 10.69

Phase plots for (a) $(j\omega\tau + 1)^{+1}$ and (b) $(j\omega\tau + 1)^{-1}$.

**FIGURE 10.70**

Magnitude plot for a second-order term in the (a) numerator and (b) denominator.

**FIGURE 10.71**

Phase plot for a second-order term in the (a) numerator and (b) denominator.

Example 10.17: Bode Plot Sketching

Plot the Bode magnitude and phase for the system with the transfer function

$$KG(s) = K \frac{s+2}{s(s^2+8s+400)}$$

where $K = 1$.

Solution

Step 1: Convert the transfer function to the frequency response function

$$KG(j\omega) = \frac{j\omega + 2}{j\omega[(j\omega)^2 + 8j\omega + 400]} = \frac{0.005((j\omega/2) + 1)}{j\omega[(j\omega/20)^2 + 2(0.2)(j\omega/20) + 1]}$$

Note that first-order and second-order terms are expressed in their corresponding basic forms, $j\omega\tau + 1$ and $(j\omega/\omega_n)^2 + 2\zeta j\omega/\omega_n + 1$.

Step 2: Identify the basic terms and the corner frequencies associated with first-order and second-order terms.

The basic terms in this example are listed as follows:

1. One constant term 0.005
2. One integral term $1/j\omega$
3. One first-order term $j\omega/2 + 1$ in the numerator with $1/\tau = 2$ rad/s
4. One second-order term in the denominator with $\omega_n = 20$ rad/s and $\zeta = 0.2$

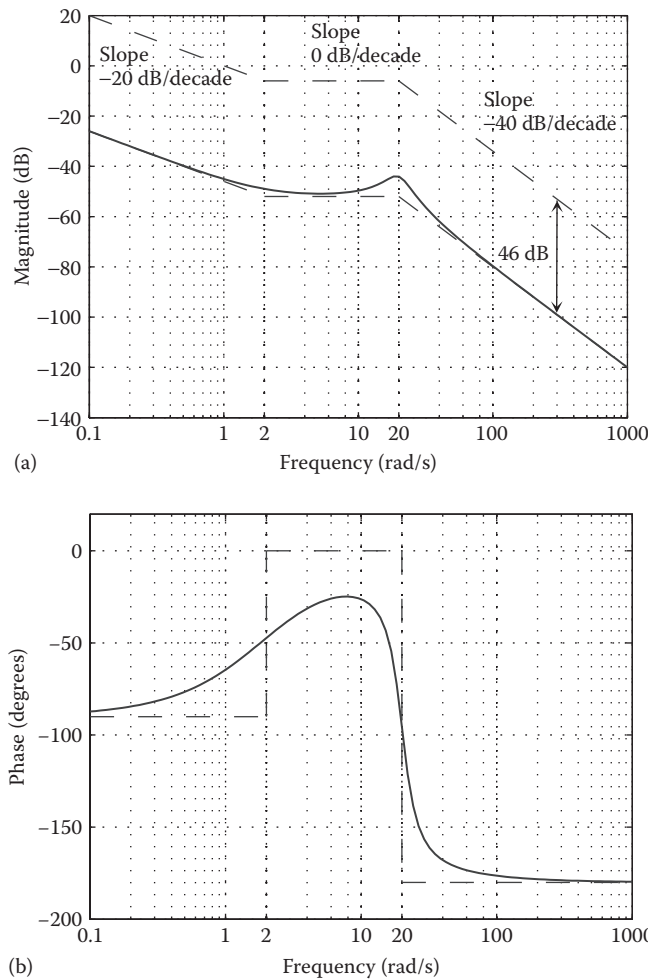
Step 3: Draw the asymptotes for the magnitude curve.

We start with indicating the corner frequencies on the frequency axis. Then, the asymptote for the derivative term is drawn through the point (1 rad/s, 0 dB) with a slope of -20 dB/decade. The asymptote is extended until the first corner frequency 2 rad/s is met, which is associated with the first-order term in the numerator. At the first corner frequency, the slope increases by 20 dB/decade and changes to 0 dB/decade. We then continue extending the asymptote until the second corner frequency 20 rad/s is met, which is associated with the second-order term in the denominator. At the second corner frequency, the slope decreases by -40 dB/decade and changes from 0 dB/decade to -40 dB/decade. Finally, we consider the effect of the constant term 0.005 by sliding the asymptotes downward by 46 dB (i.e., $20\log_{10}0.005 = -46$ dB). This completes the composite magnitude asymptotes.

The approximate Bode magnitude plot can be obtained by drawing a smooth curve following the asymptotes. The magnitude is 3 dB above the asymptote at the first-order numerator corner frequency. A resonant peak is sketched at the second-order denominator corner frequency. Note that the associated damping ratio ζ is 0.2. Thus, the peak above the asymptote is about $20\log_{10}(1/(2\zeta)) = 7.9588$ dB. [Figure 10.72a](#) shows the magnitude plot and the asymptotes.

Step 4: Draw the asymptotes for the phase curve.

Following the same procedure as in Step 3, we can sketch the asymptotes for the composite phase plot. We start with sketching the asymptote for the derivative term with a horizontal line at -90° . The phase changes by 90° at the first-order numerator corner frequency 2 rad/s, and -180° at the second-order denominator corner frequency 20 rad/s. Because the phase of a constant is 0° for any frequency, the constant is not considered when sketching the phase plot. [Figure 10.72b](#) shows the phase plot and the asymptotes.

**FIGURE 10.72**

Bode plot for the system in Example 10.17: (a) magnitude plot with asymptotes and (b) phase plot with asymptotes.

In summary, the asymptote for a composite magnitude or phase curve is plotted by starting with the $(j\omega)^{\pm n}$ term, and changing the slope or the phase at each corner frequency depending on whether the corner frequency is associated with a first-order or second-order term in the numerator or denominator. For first-order terms, the changes of slope and phase are +20 dB/decade and +90°, respectively, when in the numerator, and -20 dB/decade and -90°, respectively, when in the denominator. For second-order terms, the changes of slope and phase are +40 dB/decade and +180°, respectively, when in the numerator, and -40 dB/decade and -180°, respectively, when in the denominator. The asymptote for the magnitude is completed by shifting it up or down depending on the value of the constant term.

Bode plots can also be obtained experimentally by measuring the output amplitude and phase of a plant undergoing a sinusoidal input excitation. Following the rules of sketching Bode plots, we are able to use experimental information to identify the frequency response function or transfer function of a dynamic system.

Example 10.18: System Identification in Frequency Domain

The Bode plot of a dynamic system is shown in Figure 10.73, in which the asymptotes are also given. Following the rules of sketching Bode plots, find the transfer function of the system.

Solution

As shown in Figure 10.73, the Bode magnitude plot starts from -20 dB, the slope changes at 10 rad/s from 0 dB/decade to -40 dB/decade, and there is a lightly damped resonant peak at 10 rad/s. This implies that the frequency response function of the system consists of one constant term and one second-order term in the denominator,

$$G(j\omega) = \frac{K}{(j\omega/\omega_n)^2 + 2\zeta(j\omega/\omega_n) + 1}$$

Note that the Bode phase plot starts from 0° and changes at 10 rad/s to -180° . This also indicates a second-order term in the denominator.

The initial magnitude -20 dB $= 20 \log_{10} K$, which gives

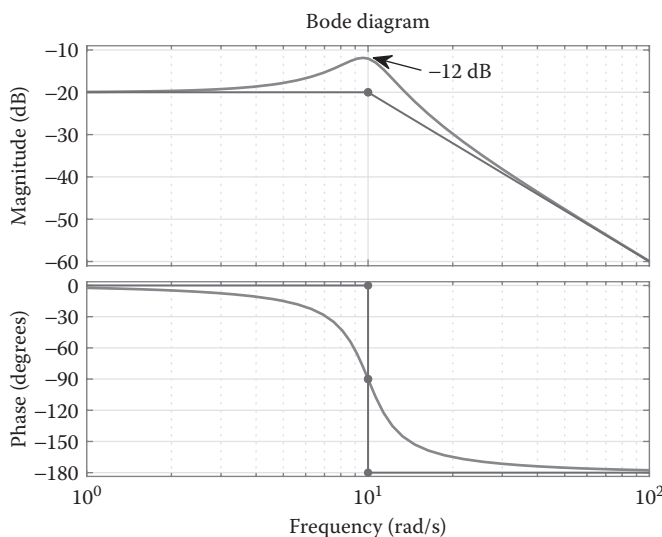
$$K = 10^{\frac{-20}{20}} = 0.1$$

The slope changes at 10 rad/s, indicating the corner frequency, also the natural frequency, of the second-order term is $\omega_n = 10$ rad/s. The value of the peak is -12 dB, or 8 dB above the asymptotes, indicating

$$20 \log_{10} \frac{1}{2\zeta} = 8 \text{ dB}$$

which gives

$$\zeta = \frac{1}{2 \left(10^{\frac{8}{20}} \right)} = 0.20$$

**FIGURE 10.73**

Bode plot for the system in Example 10.18.

Thus, the frequency response function of the system is

$$G(j\omega) = \frac{0.1}{(j\omega/10)^2 + 2(0.2)(j\omega/10) + 1}$$

and replacing $j\omega$ with s gives the transfer function of the system as

$$G(s) = \frac{10}{s^2 + 4s + 100}$$

10.6.2 Analysis Using Bode Plot

As with the root locus technique, a Bode plot can be used to determine the stability of a closed-loop system without solving for the poles. Consider the proportional feedback control system, as shown in Figure 10.65. Often, the Bode plot for $K = 1$ is first drawn. Two margins can be read from the magnitude and phase plots. As shown in Figure 10.74, the gain margin (GM) is the amount of the gain that can be added before the magnitude curve reaches 0 dB at the frequency where the phase plot crosses -180° . The phase margin (PM) is the amount of the phase that can be subtracted before the phase curve reaches -180° at the frequency where the magnitude plot crosses 0 dB.

The stability criteria are given by

$$\begin{cases} \text{GM} > 0 \text{ dB} & \text{stable} \\ \text{GM} = 0 \text{ dB} & \text{marginally stable} \\ \text{GM} < 0 \text{ dB} & \text{unstable} \end{cases} \quad (10.61)$$

or

$$\begin{cases} \text{PM} > 0^\circ & \text{stable} \\ \text{PM} = 0^\circ & \text{marginally stable} \\ \text{PM} < 0^\circ & \text{unstable} \end{cases} \quad (10.62)$$

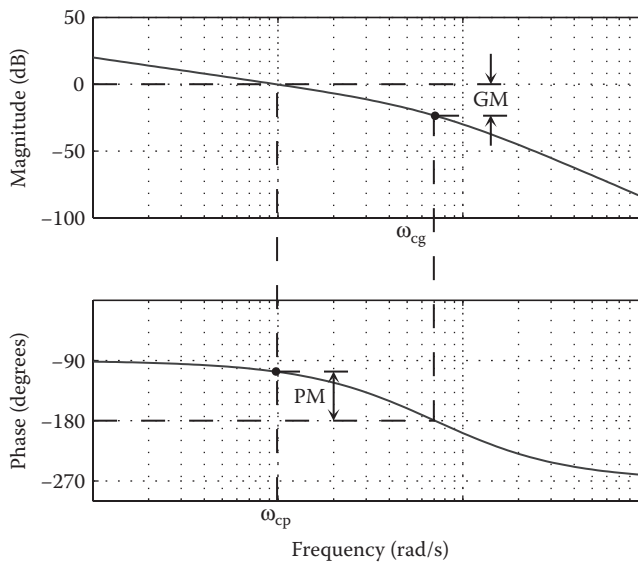


FIGURE 10.74
GM and PM.

Example 10.19: Stability Analysis Using a Bode Plot

Consider the feedback control system in [Figure 10.65](#), where

$$KG(s) = K \frac{100}{s(s+5)(s+20)}$$

and K is assumed to be 1. Plot the Bode magnitude and phase curves using MATLAB. Give comments on the stability of the closed-loop system.

Solution

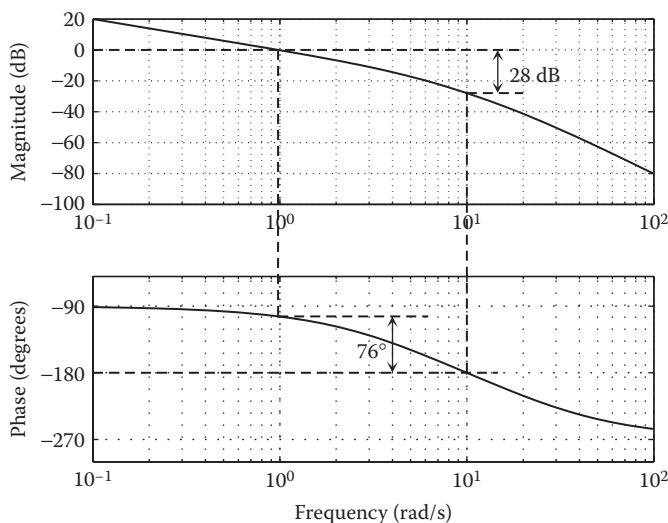
Note that the Bode plot is drawn based on the loop gain

$$L(s) = K \frac{100}{s(s+5)(s+20)}$$

The MATLAB command used to sketch the Bode plot is `bode`. Assuming $K = 1$, the following is the MATLAB session:

```
>> num = [100];
>> den = conv(conv([1 0],[1 5]),[1 20]);
>> sysL = tf(num,den);
>> bode(sysL);
```

As observed in [Figure 10.75](#), the phase curve crosses -180° at frequency 10 rad/s, and the corresponding magnitude is approximately -28 dB. This implies that a gain of 28 dB can be added before the magnitude plot reaches 0 dB at that frequency. Thus, the GM is 28 dB. The magnitude curve crosses 0 dB at frequency 1 rad/s, and the corresponding phase is -104° . This implies that a phase of 76° can be subtracted before the phase plot reaches -180° . Thus, the PM is 76° . According to Equations 10.61 and 10.62, the closed-loop system with the proportional controller $K = 1$ is stable.

**FIGURE 10.75**

Bode plot for the system in Example 10.19.

More information on stability can be extracted from the GM. Specifically, the GM when $K = 1$ gives the stability range of K for which the proportional feedback control system is stable. In our case, $GM = 28$ dB or 25, and this indicates that the closed-loop system is stable for $0 < K < 25$. We can also use the MATLAB command `margin` as follows:

```
>> [gm,pm,wcg,wcp] = margin(sysL)
```

which returns GM, PM, and the associated frequencies as defined in Figure 10.74. In our case, the GM returned by the command `margin` is 25. Note that the stability range of K can be determined in this way only for systems that change from being stable to unstable as K increases.

10.6.3 Control Design Using Bode Plot

Unlike the root locus technique, which uses time-domain performance specifications, the Bode plot technique deals with control design in the frequency domain. The requirements are defined in terms of GM, PM, bandwidth, resonant peak, and so on. If a time-domain specification is given, it will usually be converted to one in the frequency domain.

Example 10.20: Proportional Control Design Using a Bode Plot

Design a proportional controller for the cart system in Example 10.13 using the Bode plot technique.

Solution

The Bode plot for the open-loop transfer function $KG(s)$, where

$$G(s) = \frac{3.778}{s^2 + 16.883s} \quad \text{and} \quad K = 1$$

is shown in Figure 10.76.

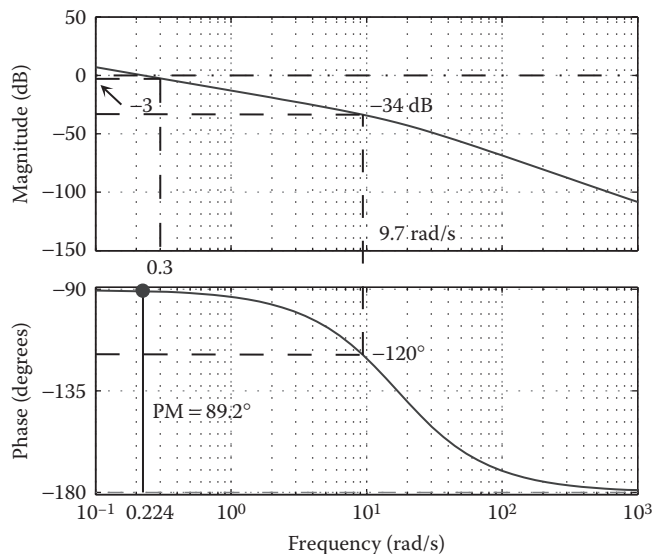


FIGURE 10.76

Bode plot for the system in Example 10.20, with $K = 1$.

Note that the requirements are given as overshoot $M_p < 10\%$ and rise time $t_r < 0.15$ s. These conditions correspond to $\zeta > 0.59$ and $\omega_n > 12.33$ rad/s. It can be shown that the relationship between the damping ratio and PM is

$$PM \approx 100\zeta \quad (10.63)$$

which, for the current example, yields the requirement $PM > 59^\circ$. In addition, the closed-loop natural frequency ω_n is related to the closed-loop bandwidth, which is somewhat greater than the frequency when the Bode magnitude plot of $KG(s)$ crosses -3 dB. Letting ω_c denote this crossover frequency, we have

$$\omega_c \leq \omega_{BW} \leq 2\omega_c \quad (10.64)$$

As the crossover frequency increases, so do the bandwidth and the natural frequency.

Figure 10.76 shows that $PM = 89.2^\circ$, which meets the requirement. However, the crossover frequency ω_c is only approximately 0.3 rad/s, which is too slow. We must adjust the value of the proportional control gain K to meet both requirements. Because the current PM is well above the requirement, we decrease the PM and pick $PM = 60^\circ$. Based on the definition of PM, this implies that the frequency at which the magnitude plot crosses 0 dB should be -120° . It is observed from Figure 10.76 that the frequency corresponding to -120° is 9.7 rad/s, where the magnitude is -34 dB. To make the magnitude 0 dB, the magnitude plot should slide upward by 34 dB. This is the effect of multiplying a constant term of

$$10^{34/20} = 50$$

which is the value of the proportional control gain K .

Let us set K to be 50, which is also what was found in Example 10.16 using the root locus design method. The Bode plot of the open-loop transfer function $KG(s)$ with the new value of K is shown in Figure 10.77. The PM is 60.1° and the crossover frequency ω_c is 12.6 rad/s. The Bode plot with $K = 1$ is also shown in Figure 10.77. Comparing the two magnitude plots, we find that the magnitude corresponding to $K = 50$ is 34 dB above the one corresponding to $K = 1$, as designed.

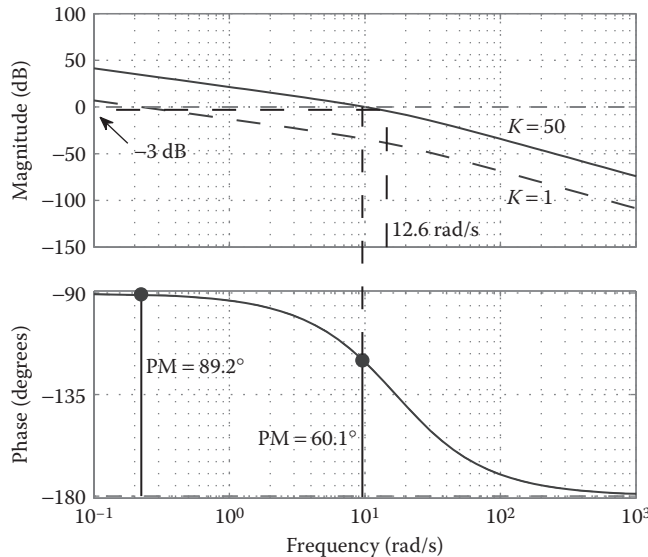


FIGURE 10.77

Bode plots for the system in Example 10.20, with $K = 50$ and $K = 1$.

PROBLEM SET 10.6

1. Sketch the asymptotes of the Bode plot magnitude and phase for the following open-loop transfer functions. Make sure to give the corner frequencies, slopes of the magnitude plot, and phase angles. Verify the results using MATLAB.

a. $G(s) = \frac{s+1}{s+10}$

b. $G(s) = \frac{s+1}{(s+10)(s+1000)}$

c. $G(s) = \frac{s+1}{s(s+10)(s+1000)}$

d. $G(s) = \frac{(s+1)^2}{s(s+10)(s+1000)}$

2. Repeat Problem 1 for the following open-loop transfer functions.

a. $G(s) = \frac{1}{s^2 + 4s + 100}$

b. $G(s) = \frac{s+0.5}{s^2 + s + 25}$

c. $G(s) = \frac{s^2 + 0.1s + 25}{s^2 + 0.24s + 144}$

d. $G(s) = \frac{100(s^2 + 7s + 49)}{(s+1)(s+500)}$

3.  For each of the following open-loop transfer functions, construct a Bode plot for $K = 1$ using the MATLAB command `bode`. Estimate the GM, PM, and their associated crossover frequencies from the plot. Verify the results using the MATLAB command `margin`. Determine the stability of the corresponding closed-loop system.

a. $KG(s) = \frac{K}{s(s^2 + 2s + 64)}$

b. $KG(s) = K \frac{1000}{(s+8)(s^2 + 2s + 4)}$

c. $KG(s) = K \frac{s+0.1}{s(s+5)^2}$

d. $KG(s) = K \frac{1}{(s+5)(s+0.5)^2}$

4. The Bode plot of a dynamic system is shown in [Figure 10.78](#), in which the asymptotes are also given. Following the rules of sketching Bode plots, find the transfer function of the system.
5. [Figure 10.79](#) shows the Bode plot for an open-loop transfer function $KG(s)$ with $K = 500$.
- Determine the stability of the closed-loop system with $K = 500$.
 - Determine the value of K that would yield a PM of 45° .

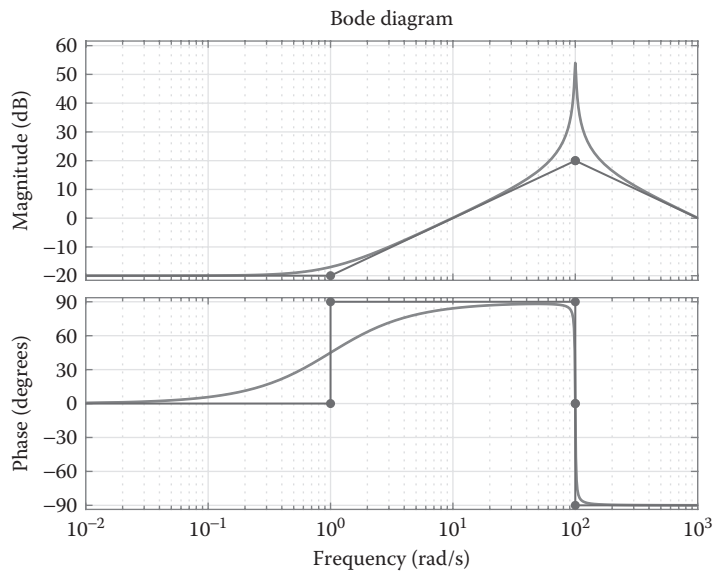


FIGURE 10.78
Problem 4.

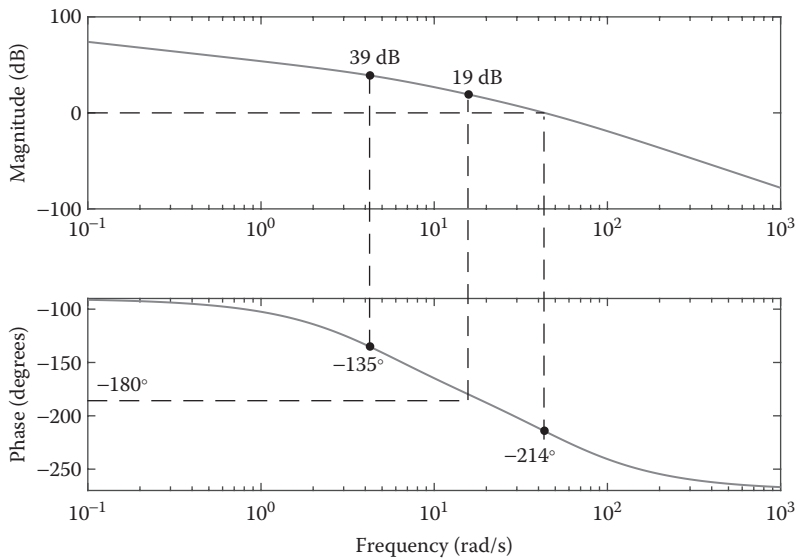


FIGURE 10.79
Problem 5.

6. The Bode plot for an open-loop transfer function $KG(s)$ is shown in [Figure 10.80](#).
 - a. Determine the stability of the closed-loop system.
 - b. Assume that the proportional control gain K is increased by a factor of 10 and 100, respectively. Will the closed-loop system still be stable with the new values of K ?

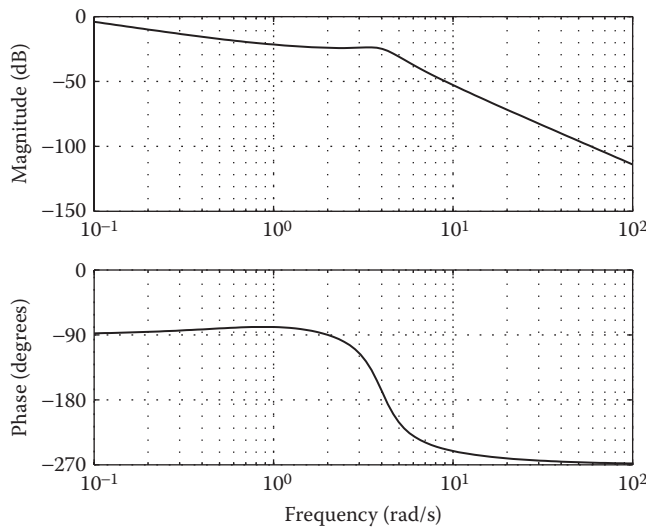


FIGURE 10.80
Problem 6.

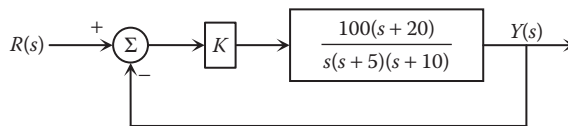


FIGURE 10.81
Problem 7.

7. Consider the unity negative feedback system shown in [Figure 10.81](#).
 - a. Use MATLAB to obtain the Bode plot $KG(s)$ for $K = 5$.
 - b. Determine the stability of the closed-loop system when $K = 5$ using the stability margins.
 - c. Determine the value of K that would yield a PM of 40° .
 - d. Verify the result obtained in Part (c) by using the MATLAB command `margin`.
8. Reconsider the feedback system in [Figure 10.64](#). Using the Bode plot technique, find a value of K such that the maximum overshoot in the response to a unit-step reference input is less than 20% and the 2% settling time is less than 1.1 s. Plot the unit-step response of the closed-loop system to verify the result.

10.7 Full-State Feedback

Unlike the root locus and Bode plot techniques, the state-space method works directly with mathematical models in state-space form instead of transfer function form. Often, use of the state-space method is referred to as modern control design, and use of transfer-function-based methods, such as the root locus and Bode plot, is referred to as classical

control design. Compared with the techniques based on transfer functions, the state-space method provides a convenient and compact way to model and analyze systems with multiple inputs and multiple outputs. This is one of the advantages of state-space design because most practical systems have more than one control input or more than one measured output. In this section, we only discuss single-input–single-output systems to show the basic ideas of state-space design. We first show how to analyze stability of a system whose model is given in state-space form. Two other important properties of control systems, known as controllability and observability, are also briefly introduced. Then, we will learn how to design a full-state feedback controller using the pole placement method.

10.7.1 Analysis of State-Space Equations

Consider a linear dynamic system with single input, single output, and n states. The state-space representation is written in the form (see more details in [Section 4.2](#))

$$\begin{aligned}\dot{\mathbf{x}} &= \mathbf{A}\mathbf{x} + \mathbf{B}u \\ y &= \mathbf{C}\mathbf{x} + Du\end{aligned}\tag{10.65}$$

where u , y , and D are scalars. The stability characteristics of a dynamic system in state-space form can be determined by the eigenvalues of matrix $\mathbf{A}$, which are the roots of

$$|s\mathbf{I} - \mathbf{A}| = 0\tag{10.66}$$

known as the characteristic equation (see more details in [Section 3.3](#)).

Example 10.21: Stability Analysis in State Space

- a. Compute the poles of the system described by

$$\begin{aligned}\dot{\mathbf{x}} &= \begin{bmatrix} 0 & 1 \\ 0 & -16.883 \end{bmatrix} \mathbf{x} + \begin{bmatrix} 0 \\ 3.778 \end{bmatrix} u \\ y &= \begin{bmatrix} 1 & 0 \end{bmatrix} \mathbf{x}\end{aligned}$$

- b. Verify the results by converting the state-space representation to a transfer function and then identifying the poles of the transfer function.

Solution

- a. The characteristic equation is

$$|s\mathbf{I} - \mathbf{A}| = \begin{vmatrix} s & -1 \\ 0 & s + 16.883 \end{vmatrix} = s^2 + 16.883s = 0$$

which yields the poles $s_1 = 0$ and $s_2 = -16.883$.

- b. As presented in [Section 4.4](#), state-space equations for a single-input–single-output system can be converted to a transfer function using

$$G(s) = \mathbf{C}(s\mathbf{I} - \mathbf{A})^{-1}\mathbf{B} + D$$

Substituting the system matrices $\mathbf{A}$, $\mathbf{B}$, $\mathbf{C}$, and D , which has the value of 0 in this example, gives

$$\begin{aligned}
 G(s) &= \begin{bmatrix} 1 & 0 \end{bmatrix} \begin{bmatrix} s & -1 \\ 0 & s+16.883 \end{bmatrix}^{-1} \begin{bmatrix} 0 \\ 3.778 \end{bmatrix} + 0 = \begin{bmatrix} 1 & 0 \end{bmatrix} \frac{\begin{bmatrix} s+16.883 & 1 \\ 0 & s \end{bmatrix}}{s^2+16.883s} \begin{bmatrix} 0 \\ 3.778 \end{bmatrix} \\
 &= \frac{3.778}{s^2+16.883s}
 \end{aligned}$$

The characteristic equation is $s^2 + 16.883s = 0$, which yields the poles at 0 and -16.883 . The results agree with the poles obtained in Part (a).

Two other important properties for a control system are controllability and observability. Before we introduce their definitions, let us consider the following example.

Example 10.22: Controllability and Observability

Consider a dynamic system given by $G(s) = 2/(s+4)$, which can be written in the state-space form

$$\begin{aligned}
 \dot{x}_1 &= -4x_1 + 2u \\
 y &= x_1
 \end{aligned}$$

- a. A new state is added and the resulting state-space equation is

$$\begin{aligned}
 \dot{x}_1 &= -4x_1 + 2u \\
 \dot{x}_2 &= -x_2 \\
 y &= x_1 + 3x_2
 \end{aligned}$$

Determine the transfer function for this new model.

- b. Determine the transfer function for another model with state-space form

$$\begin{aligned}
 \dot{x}_1 &= -4x_1 + 2u \\
 \dot{x}_2 &= -x_2 + u \\
 y &= x_1
 \end{aligned}$$

Solution

- a. The system matrices **A**, **B**, **C**, and **D** are

$$\mathbf{A} = \begin{bmatrix} -4 & 0 \\ 0 & -1 \end{bmatrix}, \quad \mathbf{B} = \begin{bmatrix} 2 \\ 0 \end{bmatrix}, \quad \mathbf{C} = \begin{bmatrix} 1 & 3 \end{bmatrix}, \quad D = 0$$

The transfer function is

$$G(s) = \begin{bmatrix} 1 & 3 \end{bmatrix} \begin{bmatrix} s+4 & 0 \\ 0 & s+1 \end{bmatrix}^{-1} \begin{bmatrix} 2 \\ 0 \end{bmatrix} + 0 = \frac{\begin{bmatrix} 1 & 3 \end{bmatrix} \begin{bmatrix} s+1 & 0 \\ 0 & s+4 \end{bmatrix} \begin{bmatrix} 2 \\ 0 \end{bmatrix}}{(s+1)(s+4)} = \frac{2}{s+4}$$

- b. Similarly, we have

$$\mathbf{A} = \begin{bmatrix} -4 & 0 \\ 0 & -1 \end{bmatrix}, \quad \mathbf{B} = \begin{bmatrix} 2 \\ 1 \end{bmatrix}, \quad \mathbf{C} = \begin{bmatrix} 1 & 0 \end{bmatrix}, \quad D = 0$$

and

$$G(s) = \begin{bmatrix} 1 & 0 \end{bmatrix} \begin{bmatrix} s+4 & 0 \\ 0 & s+1 \end{bmatrix}^{-1} \begin{bmatrix} 2 \\ 1 \end{bmatrix} + 0 = \frac{\begin{bmatrix} 1 & 0 \end{bmatrix} \begin{bmatrix} s+1 & 0 \\ 0 & s+4 \end{bmatrix} \begin{bmatrix} 2 \\ 1 \end{bmatrix}}{(s+1)(s+4)} = \frac{2}{s+4}$$

Note that the dynamic models in Parts (a) and (b) are second-order systems, with state-space forms that differ from the original first-order system. However, they both end up with the same transfer function as the given first-order system due to *pole-zero cancellation*. As seen in Part (a), the second state cannot be affected by the input matrix $\mathbf{B}$,

$$G(s) = \frac{\begin{bmatrix} s+1 & 3(s+4) \end{bmatrix}}{(s+1)(s+4)} \begin{bmatrix} 2 \\ 0 \end{bmatrix}$$

This implies that the second state is uncontrollable by the actuator defined by matrix $\mathbf{B}$. Similarly, in Part (b), the second state cannot be seen by the output matrix $\mathbf{C}$,

$$G(s) = \begin{bmatrix} 1 & 0 \end{bmatrix} \frac{\begin{bmatrix} 2(s+1) \\ s+4 \end{bmatrix}}{(s+1)(s+4)}$$

This implies that the second state is unobservable by the sensor defined by matrix $\mathbf{C}$.

Now we introduce more rigorous definitions of controllability and observability. A system is controllable if there exists a control signal $u(t)$ that will take the state of the system from any initial state $\mathbf{x}_0$ to any desired final state $\mathbf{x}_f$ in a finite amount of time. A system is observable if for any initial state $\mathbf{x}_0$, there is a finite time τ such that $\mathbf{x}_0$ can be determined from $u(t)$ and $y(t)$ for $0 \leq t \leq \tau$.

An n th-order single-input-single-output system is controllable if and only if the square matrix given by

$$\mathbf{P} = \begin{bmatrix} \mathbf{B} & \mathbf{AB} & \mathbf{A}^2\mathbf{B} & \cdots & \mathbf{A}^{n-1}\mathbf{B} \end{bmatrix} \quad (10.67)$$

is nonsingular, where $\mathbf{P}$ is called the controllability matrix. Similarly, the system is observable if and only if the square matrix given by

$$\mathbf{Q} = \begin{bmatrix} \mathbf{C} \\ \mathbf{CA} \\ \mathbf{CA}^2 \\ \vdots \\ \mathbf{CA}^{n-1} \end{bmatrix} \quad (10.68)$$

is nonsingular, where $\mathbf{Q}$ is called the observability matrix. Because a nonsingular square matrix is of full rank (Section 3.1), we can also check the rank of matrix $\mathbf{P}$ or $\mathbf{Q}$ to determine the controllability or observability.

Example 10.23: Controllability and Observability

Determine the controllability and observability of the two second-order systems given in Example 10.22.

Solution

For the system in Part (a), the controllability matrix is

$$\mathbf{P} = [\mathbf{B} \quad \mathbf{AB}] = \begin{bmatrix} 2 & -8 \\ 0 & 0 \end{bmatrix}$$

which is singular. Thus, the system is uncontrollable. The observability matrix is

$$\mathbf{Q} = \begin{bmatrix} \mathbf{C} \\ \mathbf{CA} \end{bmatrix} = \begin{bmatrix} 1 & 3 \\ -4 & -3 \end{bmatrix}$$

which is nonsingular. Thus, the system is observable.

For the system in Part (b), the controllability matrix is

$$\mathbf{P} = [\mathbf{B} \quad \mathbf{AB}] = \begin{bmatrix} 2 & -8 \\ 1 & -1 \end{bmatrix}$$

which is nonsingular. Thus, the system is controllable. The observability matrix is

$$\mathbf{Q} = \begin{bmatrix} \mathbf{C} \\ \mathbf{CA} \end{bmatrix} = \begin{bmatrix} 1 & 0 \\ -4 & 0 \end{bmatrix}$$

which is singular. Thus, the system is unobservable.

Note that the two systems in Example 10.22 have the same transfer function representation but different controllability and observability properties. This implies that controllability and observability are functions of the state of the system and cannot be determined from a transfer function.

10.7.2 Control Design for Full-State Feedback

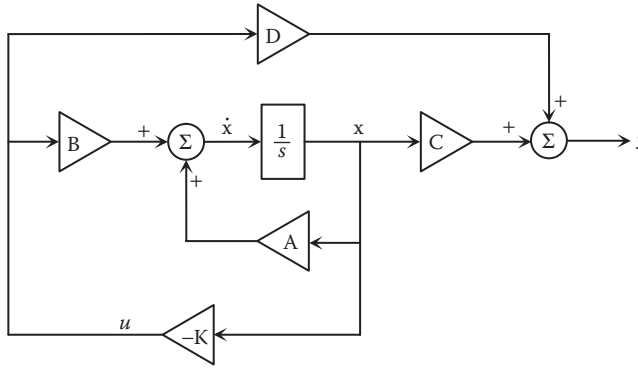
Consider an n th-order dynamic system given by Equation 10.65. If all the states are measurable, then they can be fed back and used for computing the control input

$$u = -\mathbf{K}\mathbf{x} = -\begin{bmatrix} k_1 & k_2 & \cdots & k_n \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \\ \vdots \\ x_n \end{Bmatrix} \quad (10.69)$$

where $\mathbf{K}$ is the feedback gain matrix. The control law defined by Equation 10.69 is called full-state feedback. [Figure 10.82](#) shows the block diagram of a closed-loop system with full-state feedback.

Substituting Equation 10.69 into Equation 10.65 gives the state equation of the closed-loop system, that is,

$$\dot{\mathbf{x}} = (\mathbf{A} - \mathbf{BK})\mathbf{x} \quad (10.70)$$

**FIGURE 10.82**

Block diagram of a closed-loop system with full-state feedback.

Note that the closed-loop poles are the eigenvalues of the matrix $\mathbf{A} - \mathbf{BK}$, and the closed-loop characteristic equation is

$$|s\mathbf{I} - (\mathbf{A} - \mathbf{BK})| = 0 \quad (10.71)$$

where the left-hand side is an n th-order polynomial in s containing the gains $k_1, k_2, \dots, k_n$. These gains, and hence the feedback gain matrix $\mathbf{K}$, can be determined using the pole placement method.

As discussed in [Section 10.2](#), the performance of a controlled system is associated with the closed-loop poles. If a feedback gain matrix $\mathbf{K}$ is determined based on desired pole locations, then the closed-loop system with the feedback control law $u = -\mathbf{K}\mathbf{x}$ will achieve the desired performance. This is the basic idea of pole placement. Assume that the desired locations of the closed-loop poles are $s_1, s_2, \dots$, and s_n . Note that poles of a system are the roots of the characteristic equation of the system. Thus, the desired characteristic equation is

$$(s - s_1)(s - s_2) \dots (s - s_n) = 0 \quad (10.72)$$

which is essentially the same as the closed-loop characteristic equation given by Equation 10.71,

$$|s\mathbf{I} - (\mathbf{A} - \mathbf{BK})| = (s - s_1)(s - s_2) \dots (s - s_n) \quad (10.73)$$

Equating the coefficients of like powers of s on both sides yields the values of the gains $k_1, k_2, \dots$, and k_n .

Example 10.24: Full-State Feedback Control Design

Consider the DC motor-driven cart discussed in Example 10.13, in which a PD controller was designed to satisfy the requirements: overshoot $M_p < 10\%$ and rise time $t_r < 0.15$ s.

- Find a full-state feedback controller such that the closed-loop system meets the same requirements.
- Use MATLAB to find the control gain matrix $\mathbf{K}$.

Solution

- a. The transfer function of the cart is given by

$$G(s) = \frac{Y(s)}{U(s)} = \frac{3.778}{s^2 + 16.883s}$$

where:

the output y is the position of the cart

the input u is the voltage applied to the DC motor

Using the position and the velocity as the state variables, that is, $x_1 = y$ and $x_2 = \dot{y}$, we find the state-space model as

$$\begin{aligned}\dot{\mathbf{x}} &= \begin{bmatrix} 0 & 1 \\ 0 & -16.883 \end{bmatrix} \mathbf{x} + \begin{bmatrix} 0 \\ 3.778 \end{bmatrix} u \\ y &= \begin{bmatrix} 1 & 0 \end{bmatrix} \mathbf{x}\end{aligned}$$

For a second-order system, the gain matrix $\mathbf{K}$ is a 1×2 matrix in the form

$$\mathbf{K} = \begin{bmatrix} k_1 & k_2 \end{bmatrix}$$

Applying Equation 10.71 yields the theoretical characteristic polynomial

$$\begin{aligned} |s\mathbf{I} - \mathbf{A} + \mathbf{BK}| &= \left| \begin{bmatrix} s & 0 \\ 0 & s \end{bmatrix} - \begin{bmatrix} 0 & 1 \\ 0 & -16.883 \end{bmatrix} + \begin{bmatrix} 0 \\ 3.778 \end{bmatrix} \begin{bmatrix} k_1 & k_2 \end{bmatrix} \right| \\ &= \begin{vmatrix} s & -1 \\ 3.778k_1 & s + 16.883 + 3.778k_2 \end{vmatrix} \\ &= s^2 + (16.883 + 3.778k_2)s + 3.778k_1 \end{aligned}$$

The time-domain specifications, $M_p < 10\%$ and $t_r < 0.15$ s indicate that $\zeta > 0.59$ and $\omega_n > 12.33$ rad/s. Choose the same values of ω_n and ζ as in Example 10.13, where $\omega_n = 13.5$ rad/s and $\zeta = 0.65$. The desired closed-loop poles are then located at

$$p_{1,2} = -\zeta\omega_n \pm j\omega_n\sqrt{1-\zeta^2} = -8.775 \pm 10.26j$$

Applying Equation 10.72 gives the desired characteristic polynomial

$$(s - p_1)(s - p_2) = s^2 + 2\zeta\omega_n s + \omega_n^2 = s^2 + 17.55s + 182.25$$

Equating the two characteristic polynomials, we have

$$16.883 + 3.778k_2 = 17.55$$

$$3.778k_1 = 182.25$$

which gives $\mathbf{K} = [48.24 \quad 0.18]$. So, the full-state feedback controller is

$$u = -\mathbf{K}\mathbf{x} = -48.24x_1 - 0.18x_2$$

- b.  The following MATLAB script computes a full-state feedback control gain matrix $\mathbf{K}$ using pole placement design.

```
>> A = [0 1; 0 -16.883];
>> B = [0; 3.778];
>> p = [-8.775+10.26j -8.775-10.26j];
>> K = place(A,B,p);
```

The command `place` returns the gain matrix $\mathbf{K}$, for which the full-state feedback $u = -\mathbf{K}\mathbf{x}$ places the closed-loop poles at the desired locations.

It is interesting to see that the values of the gains k_1 and k_2 are the same as the values of the proportional and derivative gains k_p and k_D found in Example 10.13. This is because of the way we selected the states. In this example, $u = -48.24y - 0.18\dot{y}$, and in Example 10.13, $u = 48.24(r - y) + 0.18 \, d(r - y)/dt$. If the reference signal r is zero, the two controllers will end up with the same expression. This implies that the closed-loop system with full-state feedback is a regulation system. For tracking control, the control law $u = -\mathbf{K}\mathbf{x}$ needs to be modified (more details can be found in control texts). Note that if a different set of state variables is selected, the corresponding gain matrix $\mathbf{K}$ will be different. The reader can solve Problem 5 in Problem Set 10.7 to verify this conclusion.

The full-state feedback control method requires that all state variables are measured. However, this is usually not a practical assumption. To make a full-state feedback controller practically implementable, an estimator or observer can be designed to compute an estimate of the state variables based on the measurements of the system. Then, the control law calculations are based on the estimated state rather than the actual state.

PROBLEM SET 10.7

1. For the system shown in Figure 10.83, derive the state-space equations using the state variables indicated. Make sure to give the $\mathbf{A}$, $\mathbf{B}$, $\mathbf{C}$, and $\mathbf{D}$ matrices. Also determine the poles of the system.
2. Repeat Problem 1 for the system shown in Figure 10.84.
3. Determine the controllability and observability for each of the following systems.

a.
$$\begin{Bmatrix} \dot{x}_1 \\ \dot{x}_2 \\ \dot{x}_3 \end{Bmatrix} = \begin{bmatrix} -5 & -3 & 0 \\ 2 & 0 & 0 \\ 0 & 1 & 0 \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \\ x_3 \end{Bmatrix} + \begin{bmatrix} 2 \\ 0 \\ 0 \end{bmatrix} u, \quad y = \begin{bmatrix} 0 & 1 & 6 \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \\ x_3 \end{Bmatrix}$$

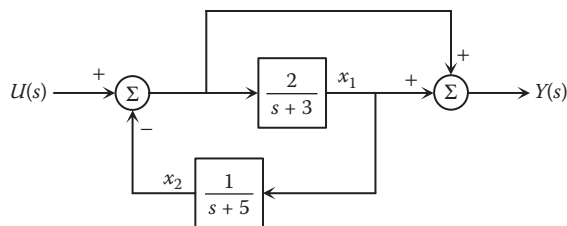
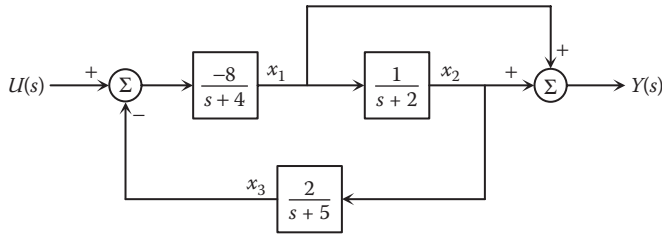


FIGURE 10.83
Problem 1.

**FIGURE 10.84**

Problem 2.

$$\text{b. } \begin{cases} \dot{x}_1 \\ \dot{x}_2 \\ \dot{x}_3 \end{cases} = \begin{bmatrix} -1 & -1 & -2 \\ 4 & 0 & 0 \\ 0 & 0 & 2 \end{bmatrix} \begin{cases} x_1 \\ x_2 \\ x_3 \end{cases} + \begin{bmatrix} 2 \\ 0 \\ 0 \end{bmatrix} u, \quad y = \begin{bmatrix} 2 & 2 & 1 \end{bmatrix} \begin{cases} x_1 \\ x_2 \\ x_3 \end{cases}$$

$$\text{c. } \begin{cases} \dot{x}_1 \\ \dot{x}_2 \\ \dot{x}_3 \end{cases} = \begin{bmatrix} 1 & 0 & 0 \\ 0 & -2 & 0 \\ 0 & 7 & -6 \end{bmatrix} \begin{cases} x_1 \\ x_2 \\ x_3 \end{cases} + \begin{bmatrix} 1 \\ 1 \\ 1 \end{bmatrix} u, \quad y = \begin{bmatrix} 1 & 1 & 0 \end{bmatrix} \begin{cases} x_1 \\ x_2 \\ x_3 \end{cases}$$

$$\text{d. } \begin{cases} \dot{x}_1 \\ \dot{x}_2 \\ \dot{x}_3 \end{cases} = \begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 4 & 3 \end{bmatrix} \begin{cases} x_1 \\ x_2 \\ x_3 \end{cases} + \begin{bmatrix} 0 \\ 0 \\ 1 \end{bmatrix} u, \quad y = \begin{bmatrix} 1 & 1 & 1 \end{bmatrix} \begin{cases} x_1 \\ x_2 \\ x_3 \end{cases}$$

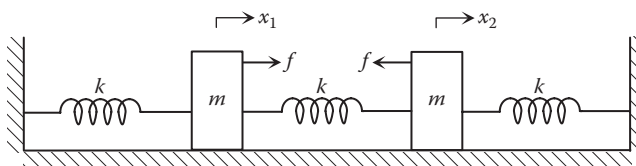
4. Consider the two-degree-of-freedom mass–spring system as shown in [Figure 10.85](#), in which two masses are to be controlled by two equal and opposite forces f . The equations of motion of the system are derived as

$$m\ddot{x}_1 + 2kx_1 - kx_2 = f$$

$$m\ddot{x}_2 - kx_1 + 2kx_2 = -f$$

Show that the system is uncontrollable. Using the concept of mode discussed in [Section 9.4](#), associate a physical meaning with the controllable and uncontrollable modes.

5. Reconsider Example 10.24. Using the approach in [Section 4.4](#), find the controllable canonical form for the plant transfer function and then design a full-state feedback controller that places the closed-loop poles at the same locations as in the example.

**FIGURE 10.85**

Problem 4.

6. A regulation system has a plant with the transfer function

$$G(s) = \frac{Y(s)}{U(s)} = \frac{5}{s^3 + 3s^2 + 4s + 6}$$

- a. Transform the plant transfer function into the state-space form with the state vector $\mathbf{x} = [y \quad \dot{y} \quad \ddot{y}]^T$.
 - b. Determine the state-feedback gain matrix $\mathbf{K}$ such that the closed-loop poles are located at $p_{1,2} = -3 \pm 4j$ and $p_3 = -8$.
 - c.  Verify the result in Part (b) by using the MATLAB command `place`.
7. Consider the system

$$\begin{Bmatrix} \dot{x}_1 \\ \dot{x}_2 \end{Bmatrix} = \begin{bmatrix} 0 & 1 \\ 0 & -1 \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix} + \begin{bmatrix} 0 \\ 10 \end{bmatrix} u, \quad y = \begin{bmatrix} 1 & 0 \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix}$$

- a. Design a state-feedback controller, so that the closed-loop poles have a damping ratio $\zeta = 0.7$ and a natural frequency $\omega_n = 3$ rad/s.
 - b.  Verify the result in Part (a) by using the MATLAB command `place`.
8. Consider the system

$$G(s) = \frac{Y(s)}{U(s)} = \frac{1}{(s+1)(s+2)(s+3)}$$

- a. Design a state-feedback controller, so that the closed-loop response has an overshoot of less than 5% and a rise time under 0.5 s. Set one of the closed-loop poles at -10 .
- b.  Verify the result in Part (a) by using the MATLAB command `place`.

10.8 Integration of Simulink and Simscape into Control Design

The root locus, Bode plot, and state-space techniques are all model-based control design methods. The main steps in a model-based control design method are plant modeling, controller analysis and synthesis, computer simulation, and real-time implementation. To control a dynamic system, a mathematical model is first derived by applying physical laws (such as Newton's second law, Kirchhoff's law, conservation of mass, etc.) or identified using experimental data. Then, closed-loop stability and performance requirements are determined by analyzing the dynamics of the plant, and a controller is designed based on the mathematical model of the plant to meet all requirements. Before implementing the controller on the real dynamic system, computer simulation is usually conducted to verify the closed-loop stability and performance requirements. Examples given in this section illustrate how to integrate Simulink and Simscape into control design to investigate the closed-loop system characteristics.

10.8.1 Control System Simulation Using Simulink

Simulink is a graphical tool that allows us to simulate a feedback control system. A Simulink model representing the block diagram in [Figure 10.1](#) can be constructed, in which the plant could be a simple linear time-invariant or even a nonlinear model. Compared with the Simulink examples of system modeling discussed in [Chapters 5](#) through [7](#), there is another important and necessary component in control system simulation, which is the controller. Several examples in [Sections 10.3](#) and [10.4](#) have shown how to build a Simulink block diagram of a control system, in which the plant is represented by a transfer function. In this section, we will consider simulation of a state feedback control system, in which the plant is represented in state-space form.

Example 10.25: Full-State Feedback Control of a Direct Current Motor–Driven Cart

Consider the full-state feedback control system discussed in [Example 10.24](#), in which the state-space representation of the plant is

$$\dot{\mathbf{x}} = \begin{bmatrix} 0 & 1 \\ 0 & -16.883 \end{bmatrix} \mathbf{x} + \begin{bmatrix} 0 \\ 3.778 \end{bmatrix} u$$

$$y = \begin{bmatrix} 1 & 0 \end{bmatrix} \mathbf{x}$$

and the mathematical model of the controller is

$$u = -\mathbf{K}\mathbf{x}$$

with $\mathbf{K} = [48.24 \ 0.18]$. Build a Simulink block diagram of the feedback control system and find the closed-loop response if the cart is initially 1 m away from the equilibrium position.

Solution

There are two ways to construct a Simulink block diagram of a full-state feedback control system. If we treat the state-space model as its scalar counterpart,

$$\dot{x} = ax + bu$$

$$y = cx$$

then we can build a block diagram as shown in [Figure 10.86](#), in which the state-space model is represented by using one Integrator block and three Gain blocks ($\mathbf{A}$, $\mathbf{B}$, and $\mathbf{C}$). Note that all the gains, including the control gain $\mathbf{K}$, are in matrix form.

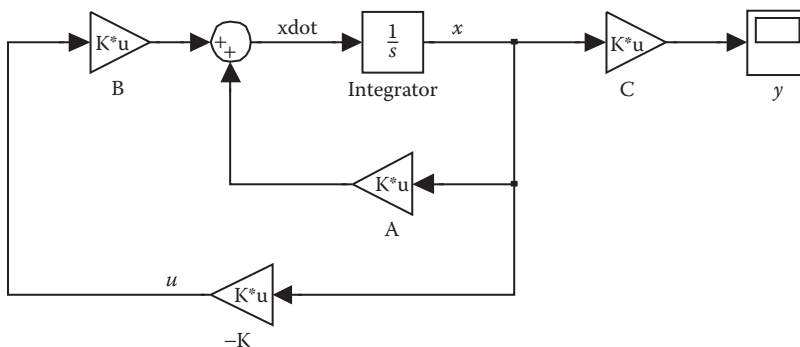
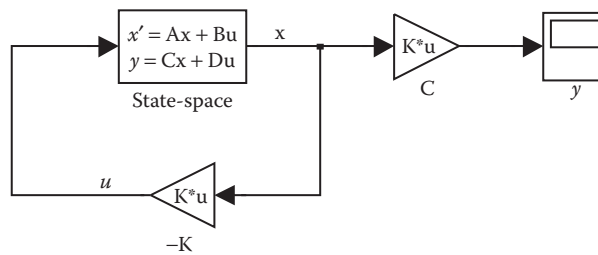


FIGURE 10.86

Simulink block diagram of a full-state feedback control system.

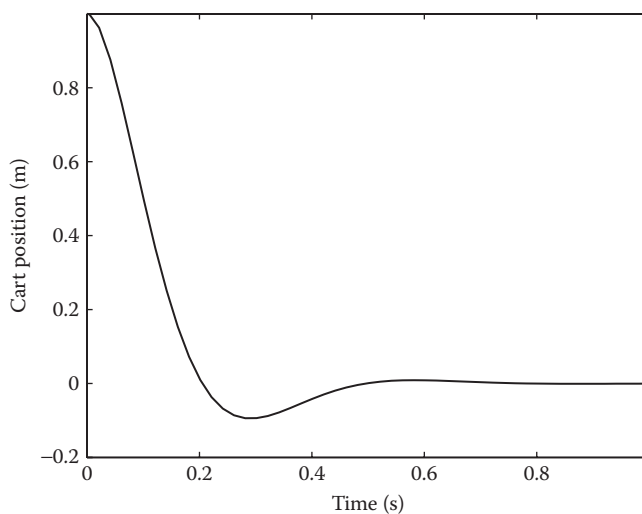
**FIGURE 10.87**

Alternative Simulink block diagram of a full-state feedback control system.

This should be specified explicitly in Simulink. Double-click each Gain block, define the corresponding matrix, and choose Matrix ($K*u$) for the Multiplication parameter. Recall that we have chosen $x_1 = y$ and $x_2 = \dot{y}$. Thus, the physical position variable is the same as the first state variable. To specify the non-zero initial position, double-click the Integrator block and type $[1; 0]$ for the Initial conditions parameter.

Figure 10.87 presents an alternative Simulink block diagram, in which the state-space model is built using the State-Space block instead of the Integrator block. Note that all state-variables must be available for a full-state feedback control system because the control signal is $u = -Kx$. To simulate this, double-click the State-Space block and define C as an identity matrix and D as a zero matrix with compatible dimensions; `eye(2)` for C and `zeros(2,1)` for D in this example. The parameter of Initial conditions has the same value as defined in Figure 10.86. To obtain the output y , a Gain block is included to define the real matrix C . It should be pointed out that the value of the Gain block corresponding to the full-state feedback controller is $-K$, not K .

Running both simulations yields the same curve, as shown in Figure 10.88, which is the resulting displacement response $y(t)$ due to the non-zero initial condition of 1 m. It is interesting to note that the curve in Figure 10.88 is a mirror image of the unit-step response curve in Figure 10.37 about the x-axis. The reason is left to the reader to find out.

**FIGURE 10.88**

Closed-loop response of the cart system with full-state feedback control.

10.8.2 Integration of Simscape into Control System Simulation

Instead of Simulink blocks, such as Transfer Fcn, State-Space, etc., Simscape models can also be integrated into control system simulation to model the open-loop plant dynamics and test closed-loop system performance.

Example 10.26: Control of a Single-Degree-of-Freedom Mass–Spring–Damper System

Consider a single-degree-of-freedom mass–spring–damper system as shown in Figure 5.29, where $m = 2$ kg, $b = 2$ N s/m, and $k = 50$ N/m. A PD controller, $f = -k_p x - k_D \dot{x}$, is designed to adjust the input force f , so that the mass block can be maintained in the equilibrium position regardless of disturbance forces applied to the block. The performance requirements of the closed-loop system are overshoot $M_p < 5\%$ and rise time $t_r < 0.25$ s.

- Design a PD controller to meet the performance requirements.
- Build a block diagram of the feedback control system, in which the plant is constructed using Simscape blocks and the controller is constructed using Simulink blocks. Find the closed-loop response if the mass block is initially 0.1 m away from the equilibrium position.

Solution

- The dynamics of the plant is described by

$$m\ddot{x} + b\dot{x} + kx = f$$

where the control force

$$f = -k_p x - k_D \dot{x}$$

Combining the two equations gives the dynamics of the closed-loop system, as

$$m\ddot{x} + (b + k_D)\dot{x} + (k + k_p)x = 0 \quad \xRightarrow{\text{Divide by } m} \quad \ddot{x} + \frac{b + k_D}{m}\dot{x} + \frac{k + k_p}{m}x = 0$$

which is a second-order system. Thus, the coefficients in the previous differential equation can be related to the natural frequency and damping ratio of the closed-loop system via

$$\begin{aligned} \frac{b + k_D}{m} &= 2\zeta\omega_n \\ \frac{k + k_p}{m} &= \omega_n^2 \end{aligned}$$

The requirement for overshoot indicates

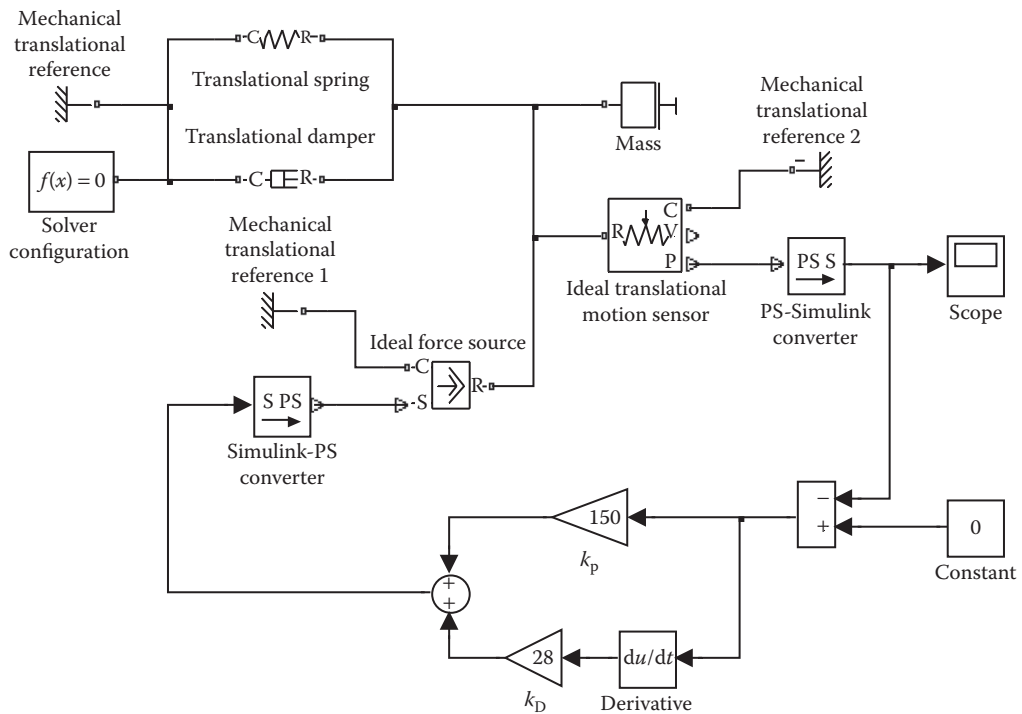
$$\zeta > 0.69$$

Choosing $\zeta = 0.75$ and inserting into the requirement for rise time gives

$$\omega_n > 9.26 \text{ rad/s}$$

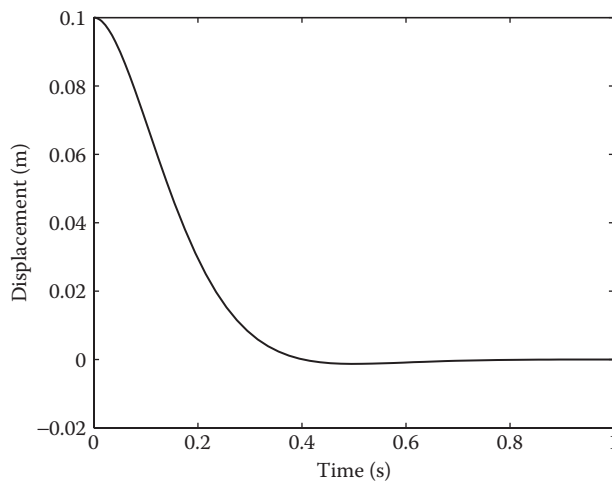
Pick $\omega_n = 10$ rad/s. Simultaneous solution of the two relations given earlier yields $k_p = 150$ and $k_D = 28$.

- Figure 10.89 is the block diagram of the resulting feedback control system built using Simulink and Simscape. The plant is constructed based on the physical mass–spring–damper system and the details on Simscape modeling can be found in Example 5.4. The controller is constructed using Simulink blocks, and its structure is similar to the PD control discussed in Section 10.4 with the reference signal r set as 0. To specify a non-zero initial position, double-click

**FIGURE 10.89**

Simscape block diagram of the feedback control system in Example 10.26.

the Translational Spring block, type 0.1 for the Initial deformation, and choose the unit as m. This implies that the spring is initially elongated by 0.1 m. Also, double-click the Ideal Translational Motion Sensor block, type 0.1 for the Initial position, and choose the unit as m. The corresponding displacement response of the system is shown in Figure 10.90.

**FIGURE 10.90**

Closed-loop response of the mass-damper-spring system with PD control.

PROBLEM SET 10.8

1.  Consider the control system shown in [Figure 10.42](#) (Problem Set 10.4, Problem 3). Using the results obtained in Part (a) of the cited problem, build a Simulink block diagram to simulate the feedback control system and find the unit-step response of the closed-loop system.
2.  Repeat Problem 1 for the control system shown in [Figure 10.43](#) (Problem Set 10.4, Problem 4).
3.  Repeat Problem 1 for the control system shown in [Figure 10.44](#) (Problem Set 10.4, Problem 5).
4.  Repeat Problem 1 for the control system shown in [Figure 10.45](#) (Problem Set 10.4, Problem 6).
5.  Consider the control system shown in [Figure 10.63](#) (Problem Set 10.5, Problem 7). Using the results obtained in Part (b) of the cited problem, build a Simulink block diagram to simulate the feedback control system and find the unit-step response of the closed-loop system.
6.  Repeat Problem 5 for the control system shown in [Figure 10.64](#) (Problem Set 10.5, Problem 8).
7.  Consider Problem 6 in Problem Set 10.7. Using the state-space model obtained in Part (a) and the full-state feedback controller obtained in Part (b), build a Simulink block diagram to simulate the resulting feedback control system. Find the closed-loop response if the initial conditions are $y(0) = 0.1$, $\dot{y}(0) = 0$, and $\ddot{y}(0) = 0$.
8.  Consider Problem 7 in Problem Set 10.7. Using the full-state feedback controller obtained in Part (b), build a Simulink block diagram to simulate the resulting feedback control system. Find the closed-loop response if the initial conditions are $x_1(0) = 0.1$ and $x_2(0) = 0$.
9. Consider the rotational mass–spring–damper system in Example 5.23. A PD controller, $\tau = -k_p\theta - k_D\dot{\theta}$, is designed to adjust the input torque τ so that the rotational disk can quickly return to the equilibrium position regardless of disturbances applied to the system. The performance requirements of the closed-loop system are overshoot $M_p < 5\%$ and rise time $t_r < 0.004$ s.
 - a. Design a PD controller to meet the performance requirements.
 - b.  Build a block diagram of the feedback control system, in which the plant is constructed using Simscape blocks and the controller is constructed using Simulink blocks. Find the closed-loop response if the disk is initially 0.1 rad away from the equilibrium position.
10.  Consider the mass–spring–damper system shown in [Figure 5.107](#) (Example 5.22). Assume that f is a control force to maintain the system at equilibrium regardless of disturbances applied to the system.
 - a. Design a full-state feedback controller such that the closed-loop poles are located at $-10 \pm 10j$, -15 , and -16 . Assume the state vector $\mathbf{x} = [x_1 \ x_2 \ \dot{x}_1 \ \dot{x}_2]^T$.
 - b. Build a block diagram of the feedback control system, where the plant is constructed using Simscape blocks and the controller is constructed using Simulink blocks. Find the closed-loop response if mass 1 is initially 0.1 m from the equilibrium position.

10.9 Summary

This chapter presented an introduction to feedback control systems. The essential components of a feedback control include a system we want to control, a controller we need to design, an actuator used to drive the controlled system, and a sensor used to measure the system output. Generally, the controlled system and the actuator are intimately connected, and they can be combined as one component called the plant. Unlike open-loop control, the output signal of the plant in a feedback control system is measured and fed back for use in computing the control signal. In contrast to open-loop control, feedback can be used to stabilize unstable systems, reduce steady-state errors to disturbances, improve reference tracking performance, and reduce sensitivity to parameter variations.

Stability and performance are two important considerations in control. A linear time-invariant system is said to be stable if and only if all its poles have negative real parts and is unstable otherwise. In terms of the pole locations in the s -plane, the imaginary axis is the stability boundary between the stable left-half s -plane and the unstable right-half s -plane. Solving for the poles of a high-order linear system by hand is not an easy task. Routh's stability criterion is a method of obtaining information about pole locations without solving for the poles. A system is stable if and only if all the elements in the first column of the Routh array are positive.

The locations of poles in the s -plane are also associated with performance measures, which are rise time t_r , overshoot M_p , peak time t_p , and settling time t_s in the time domain, and bandwidth ω_{BW} and resonant peak M_r in the frequency domain. For a second-order system with poles at $-\zeta\omega_n \pm j\omega_n\sqrt{1-\zeta^2}$, the correspondences between the system parameters and the time-domain specifications are given by

$$t_r \approx \frac{1.12 - 0.078\zeta + 2.230\zeta^2}{\omega_n}$$

$$M_p = e^{-\pi\zeta/\sqrt{1-\zeta^2}}$$

$$t_p = \frac{\pi}{\omega_d}$$

$$t_s = -\frac{\ln(\Delta\sqrt{1-\zeta^2})}{\zeta\omega_n}$$

where Δ is a small value, such as 1%, 2%, and 5%. The resonant peak M_r is similar to the overshoot M_p , both of which are related to the damping ratio ζ , whereas the bandwidth ω_{BW} is similar to the rise time t_r , both of which are related to the natural frequency ω_n .

The PID controller is a generic feedback control structure widely used in industries. It is described by the transfer function

$$\frac{U(s)}{E(s)} = k_p + \frac{k_I}{s} + k_D s$$

In general, a larger proportional gain k_p results in a faster response and a smaller steady-state error. However, an excessively large proportional gain k_p leads to lightly damped oscillations and even instability. A larger integral gain k_i reduces steady-state errors more quickly, but reduces damping leading to a larger overshoot. A larger derivative control decreases the overshoot, but slows down the speed of response. The PID gains can be tuned using the reaction curve method or the ultimate sensitivity method developed by Ziegler and Nichols.

Three different methods were introduced in this chapter for stability analysis and proportional feedback control design: root locus, Bode plot, and state-space methods. The root locus and Bode plot work with graphs obtained from open-loop transfer functions, whereas the state-space method works directly with mathematical models in state-space form.

For a negative feedback system with $KL(s)$ as the open-loop transfer function, a root locus is a graph of the closed-loop poles or the roots of the closed-loop characteristic equation

$$1 + KL(s) = 0$$

with respect to the control gain K . The rules for sketching a root locus are presented in [Section 10.5](#). Using the root locus technique, it is very easy to determine the stability of a closed-loop system when the proportional gain K varies from 0 to ∞ . For a particular value of K , the closed-loop system is stable if and only if all of the poles are in the left-half s -plane.

The Bode plot is a graph of the frequency response function, using a linear scale for magnitude (in decibels) and phase (in degrees) and a logarithmic scale for frequency (in rad/s). For a frequency response function in the form of

$$KG(j\omega) = K \frac{(j\omega - z_1)(j\omega - z_2) \cdots (j\omega - z_m)}{(j\omega - p_1)(j\omega - p_2) \cdots (j\omega - p_n)}$$

the Bode plot can be easily drawn by hand using the rules described in [Section 10.6](#). The stability of the corresponding closed-loop system can be determined by the GM and PM, both of which can be found directly by inspecting the open-loop Bode plot.

A dynamic system described in state-space form is stable if all eigenvalues of the state matrix have negative real parts. If all the states are measurable, a full-state feedback controller given by

$$u = -\mathbf{K}\mathbf{x}$$

can be designed to improve stability and performance. The feedback gain matrix $\mathbf{K}$ can be determined using the pole placement method. Closed-loop poles are selected depending on the desired transient response. Equating the theoretical and desired closed-loop characteristic polynomials

$$|s\mathbf{I} - (\mathbf{A} - \mathbf{BK})| = (s - s_1)(s - s_2) \cdots (s - s_n)$$

yields the values of the elements $k_1, k_2, \dots$, and k_n of the gain matrix $\mathbf{K}$.

Review Problems

- Consider the feedback control system as shown in Figure 10.91. Determine the range of K for closed-loop stability.
- Consider the feedback control system shown in Figure 10.92.
 - Design a PD controller such that the closed-loop poles are at $p_{1,2} = -2 \pm 2j$.
 - Estimate the rise time, overshoot, peak time, and 1% settling time for the unit-step response of the closed-loop system.
 - Use MATLAB to plot the unit-step response of the closed-loop system. Find the values of the rise time t_r , overshoot M_p , peak time t_p , and 1% settling time t_s .
- Consider the feedback control system as shown in Figure 10.93.
 - Assuming $C(s) = k_p$, determine the value of the proportional gain that makes the closed-loop system marginally stable. Find the frequency of the sustained oscillation.
 - Using the gain and the frequency obtained in Part (a), apply the ultimate sensitivity method of Ziegler–Nichols tuning rules to design a PID controller.
 - Plot the unit-step response of the resulting closed-loop system. Find the values of the rise time t_r , overshoot M_p , peak time t_p , and 1% settling time t_s .

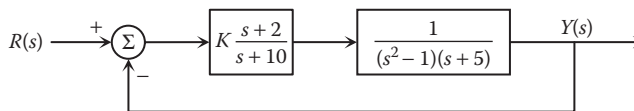


FIGURE 10.91
Problem 1.

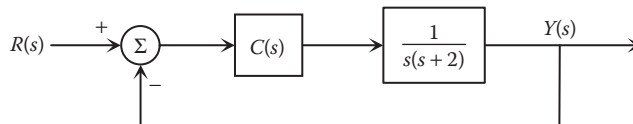


FIGURE 10.92
Problem 2.

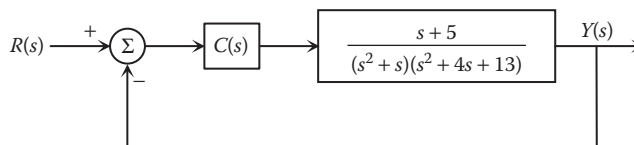


FIGURE 10.93
Problem 3.

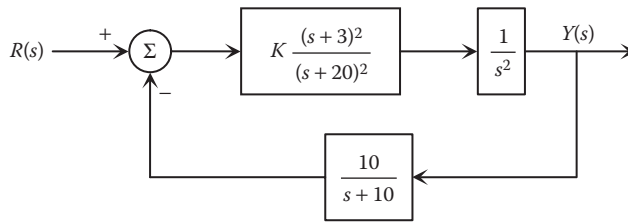


FIGURE 10.94
Problem 4.

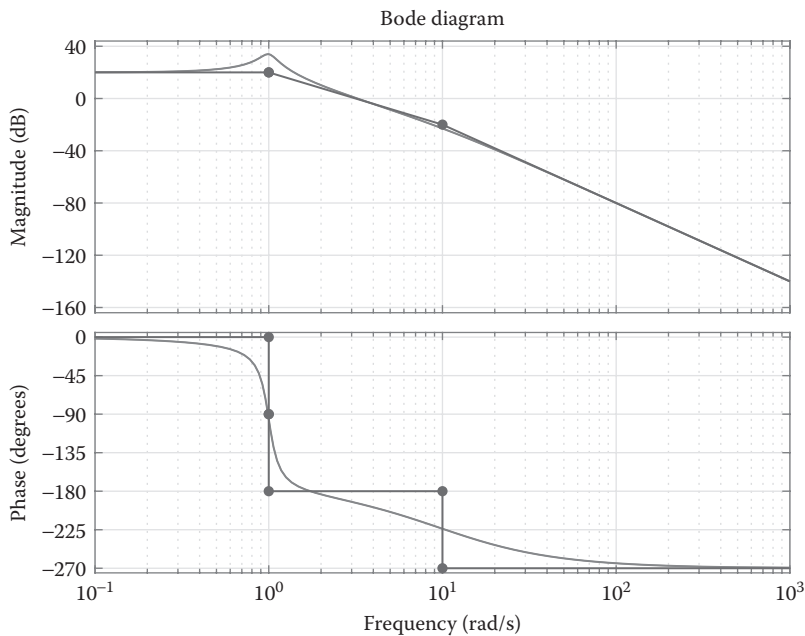
4. Consider the feedback control system as shown in [Figure 10.94](#).
 - a. Determine the value of the gain K such that the undamped natural frequency ω_n and the damping ratio ζ of the dominant closed-loop poles are roughly 2 rad/s and 0.5, respectively.
 - b. Determine the values of all closed-loop poles.
 - c. Plot the unit-step response of the resulting closed-loop system. Find the values of the rise time t_r , overshoot M_p , peak time t_p , and 2% settling time t_s .
5. Consider a unity negative feedback system with the open-loop transfer function

$$KG(s) = \frac{K}{s(s+3)(s+6)}$$

- a. Use MATLAB to draw the Bode plots for $K = 1$. Determine the range of K for which the closed-loop system will be stable.
- b. Determine the range of K for closed-loop stability by sketching the root locus.
- c. Using Routh's criterion, determine the range of K for closed-loop stability.
6. The Bode plot of a dynamic system is shown in [Figure 10.95](#), in which the asymptotes are also given. Following the rules of sketching Bode plots, find the transfer function of the system.
7. Consider the system

$$\begin{Bmatrix} \dot{x}_1 \\ \dot{x}_2 \end{Bmatrix} = \begin{bmatrix} 0 & 1 \\ 0 & -10 \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix} + \begin{bmatrix} 0 \\ 210 \end{bmatrix} u, \quad y = \begin{bmatrix} 1 & 0 \end{bmatrix} \begin{Bmatrix} x_1 \\ x_2 \end{Bmatrix}$$

- a. Design a state-feedback controller, so that the closed-loop unit-step response has an overshoot of less than 10% and a peak time under 0.5 s.
- b. Verify the results of Part (a) in MATLAB.
8. Consider the two-degree-of-freedom quarter-car model shown in [Figure 5.34](#), in which the force f , applied between the car body and the wheel–tire–axle assembly, is controlled by feedback and represents the active components of the suspension system. Assume that $f = 20568x_1 - 30493x_2 - 1278\dot{x}_1 + 3189\dot{x}_2$. Build a block diagram of the feedback control system, where the quarter-car model is constructed using Simscape blocks and the controller is constructed using Simulink blocks. Find the displacement responses $x_1(t)$ and $x_2(t)$ if initially $x_1 = -0.05$ m and $x_2 = -0.05$ m. Ignore the displacement input $z(t)$. What are the system responses $x_1(t)$ and $x_2(t)$ without control?

**FIGURE 10.95**

Problem 6.

9.  Consider the DC motor-driven wheeled mobile robot shown in Figure 6.82, in which the voltage applied to the DC motor is computed by a controller. Assume that $v_a = 2.56i - 0.37x + 4.61\dot{x} + 0.37x_r$, where x_r is a reference trajectory that the cart should follow. Build a block diagram of the feedback control system, in which the mobile robot is constructed using Simscape blocks and the controller is constructed using Simulink blocks. Find the displacement response $x(t)$ of the mobile robot if a unit-step reference command signal is sent to the system.
10. Case Study
 -  Consider the cart-inverted-pendulum system shown in Figure 5.79. Assume that the mass of the cart is 0.8 kg, the mass of the pendulum is 0.2 kg, and the length of the pendulum is 0.6 m.
 - a. Determine the poles of the linearized system. Is it stable or unstable?
 - b. Design a full-state feedback controller for the linearized system using the pole placement method. Assume that two of the closed-loop poles are complex conjugate, with a natural frequency of 3.6 rad/s and a damping ratio of 0.6. They dominate the effect of the other two poles, which are assumed at -10 and -20 .
 - c. Assume that the initial angle of the inverted pendulum is 5° away from the vertical reference line. Using the state feedback gain matrix $\mathbf{K}$ obtained in Part (b), examine the responses of the nonlinear and linearized closed-loop systems by using Simulink.

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Appendix A

TABLE A.1
The International System (SI) of Units

Basic Units		
Quantity	Unit Name	Symbol or Unit
Length	meter	m
Mass	kilogram	kg
Time	second	s
Electric current	Ampere	A
Voltage	volts	V
Temperature	kelvin	K
Amount of substance	mole	mol
Solid angle	radian	rad
SI Derived Units		
(Circular) frequency (ω)		rad/s
Acceleration (a)		m/s ²
Angular acceleration (α)		rad/s ²
Angular velocity (ω)		rad/s
Area (A)		m ²
Area moment of inertia (I)		m ⁴
Density, mass density (ρ)		kg/m ³
Dynamic viscosity (b, c)		N·s/m
Electric capacitance (C)	farad (F)	A·s/V
Electric charge (q)	coulomb (C)	A·s
Electric resistance (R)	ohm (Ω)	V/A
Force (f)	newton (N)	kg·m/s ²
Frequency	hertz (Hz)	1/s
Inductance (L)	henry (H)	V·s/A
Magnetic flux	weber (Wb)	V·s
Magnetic flux density	tesla (T)	Wb/m ²
Mass moment of inertia (I)		kg·m ²
Power	watt (W)	J/s
Pressure, mechanical stress	pascal (Pa)	N/m ²
Specific heat (c, c_p, c_v)		J/(kg·K)
Specific volume (v)		m ³ /kg
Thermal conductivity (k)		W/(s·m·K)
Velocity, speed (v)		m/s
Volume (V)		m ³
Wave length (λ)		1/m
Work (W), energy (E), heat (Q)	joule (J)	N·m

TABLE A.2

Conversion Factors

Density	$1 \text{ g/cm}^3 = 62.43 \text{ lb}_m/\text{ft}^3$
Energy	$1 \text{ cal} = 4.184 \text{ J}$
Force	$1 \text{ lb}_f = 4.45 \text{ N}$
Length	$1 \text{ in.} = 2.54 \text{ cm}$
	$1 \text{ ft} = 0.3048 \text{ m}$
	$1 \text{ mi.} = 1,609 \text{ m} = 5,280 \text{ ft}$
Mass	$1 \text{ lb}_m = 0.4536 \text{ kg} = 16 \text{ oz}$
	$1 \text{ slug} = 32.174 \text{ lb}_m$
	$1 \text{ ton} = 2,000 \text{ lb}_m$
Power	$1 \text{ W} = 3.413 \text{ Btu/hr}$
Pressure	$1 \text{ atm} = 1.0132 \times 10^5 \text{ Pa}$
Temperature	$^{\circ}\text{C} = (^{\circ}\text{F} - 32)/1.8$
	$^{\circ}\text{F} = ^{\circ}\text{C}(1.8) + 32$
	$^{\circ}\text{K} = ^{\circ}\text{C} + 273.16$
	$^{\circ}\text{R} = ^{\circ}\text{F} + 459.69$
	$1 \text{ K} = 1.8^{\circ}\text{R}$
Thermal conductivity	$1 \text{ W}/(\text{m}\cdot^{\circ}\text{C}) = 0.5778 \text{ Btu}/(\text{hr}\cdot\text{ft}\cdot^{\circ}\text{F})$
Volume	$1 \text{ liter (L)} = 1000 \text{ cm}^3 = 0.0353 \text{ ft}^3 = 1.0564 \text{ quart}$
	$1 \text{ ft}^3 = 28.316 \text{ L}$
	$1 \text{ gal} = 3.785 \text{ L} = 4 \text{ quarts}$
	$1 \text{ quart} = 2 \text{ pints} = 67.2 \text{ in.}^3 = 0.9466 \text{ L}$
	$1 \text{ pint} = 16 \text{ oz}$

Appendix B: Useful Formulas

Trigonometric Expansions

Sum to Product	Product to Sum
$\sin(a \pm b) = \sin a \cos b \pm \cos a \sin b$	$\sin a \sin b = \frac{1}{2} [\cos(a - b) - \cos(a + b)]$
$\cos(a \pm b) = \cos a \cos b \mp \sin a \sin b$	$\cos a \cos b = \frac{1}{2} [\cos(a + b) + \cos(a - b)]$
$\tan(a \pm b) = \frac{\tan a \pm \tan b}{1 \mp \tan a \tan b}$	$\sin a \cos b = \frac{1}{2} [\sin(a + b) + \sin(a - b)]$
	$\cos a \sin b = \frac{1}{2} [\sin(a + b) - \sin(a - b)]$

Double-Angle and Half-Angle Formulas

Double-Angle Formulas	Half-Angle Formulas
$\sin^2 a = \frac{1}{2} (1 - \cos 2a)$	$\cos \frac{1}{2} a = \sqrt{\frac{1}{2} (1 + \cos a)}$
$\cos^2 a = \frac{1}{2} (1 + \cos 2a)$	$\sin \frac{1}{2} a = \sqrt{\frac{1}{2} (1 - \cos a)}$
$\sin 2a = 2 \sin a \cos a$	
$\cos 2a = 2 \cos^2 a - 1$	

Hyperbolic Functions

$\sinh a = \frac{1}{2} (e^a - e^{-a})$	$\sinh(a \pm b) = \sinh a \cosh b \pm \cosh a \sinh b$
$\cosh a = \frac{1}{2} (e^a + e^{-a})$	$\cosh(a \pm b) = \cosh a \cosh b \pm \sinh a \sinh b$
$\cosh^2 a - \sinh^2 a = 1$	$\sinh^2 a = \frac{1}{2} (\cosh 2a - 1)$
	$\cosh^2 a = \frac{1}{2} (1 + \cosh 2a)$

Integration

$$\int x^n dx = \frac{x^{n+1}}{n+1} + c \quad (n \neq -1)$$

$$\int e^{ax} dx = \frac{1}{a} e^{ax} + c$$

$$\int \frac{1}{x} dx = \ln|x| + c$$

$$\int \sin x dx = -\cos x + c$$

$$\int \tan x dx = -\ln|\cos x| + c = \ln|\sec x| + c$$

$$\int \sec x dx = \ln|\sec x + \tan x| + c$$

$$\int \ln x dx = x \ln x - x + c$$

$$\int \frac{1}{a^2 + x^2} dx = \frac{1}{a} \tan^{-1} \frac{x}{a} + c$$

$$\int \frac{1}{\sqrt{a^2 - x^2}} dx = \sin^{-1} \frac{x}{a} + c$$

$$\int e^{ax} \sin bx dx = \frac{1}{a^2 + b^2} e^{ax} (a \sin bx - b \cos bx) + c$$

$$\int e^{ax} \cos bx dx = \frac{1}{a^2 + b^2} e^{ax} (a \cos bx + b \sin bx) + c$$

$$\int e^x dx = e^x + c$$

$$\int e^{g(x)} g'(x) dx = e^{g(x)} + c$$

$$\int \frac{g'(x)}{g(x)} dx = \ln|g(x)| + c$$

$$\int \cos x dx = \sin x + c$$

$$\int \cot x dx = \ln|\sin x| + c = -\ln|\csc x| + c$$

$$\int \csc x dx = \ln|\csc x - \cot x| + c$$

$$\int \frac{1}{a^2 - x^2} dx = \frac{1}{2a} \ln \left| \frac{x+a}{x-a} \right| + c = \frac{1}{a} \tanh^{-1} \frac{x}{a}$$

$$\int \frac{1}{\sqrt{a^2 + x^2}} dx = \sinh^{-1} \frac{x}{a} + c$$

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